

INTRODUCTION TO **ANTENNA** ANALYSIS USING **EM SIMULATORS**

Hiroaki Kogure, Yoshie Kogure, James C. Rautio



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Introduction to Antenna Analysis Using EM Simulators

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Introduction to Antenna Analysis Using EM Simulators

Hiroaki Kogure

Yoshie Kogure

James C. Rautio



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Preface

It is 1984. Orwell's predictions appear to be at least several decades premature. Gas costs \$1.10 per gallon. The first Apple Macintosh goes on sale. The space shuttle Discovery flies its maiden voyage. And I write a series of articles for the amateur radio magazine *QST* about antenna analysis using computers.

The most sophisticated personal computer (the term had just come into use) of the day had just been introduced, the IBM-PC, with an incredible 640 kB of RAM and a clock speed exceeding 4 MHz. That is higher in frequency than the carrier frequency of my first ham radio contact. It is an absolutely amazing time!

The program I wrote for those *QST* articles, called "Annie," is what got my company started. We sold some 250 copies worldwide. Shortly after the first article appeared, I received a letter from Aki Kogure, first author of this book. He wanted to try publicizing and selling Annie in Japan. Why not? I decided to take a chance and sent the software off to him. It was amazing. He sold another 250 copies in Japan alone!

Aki, and his wonderful wife, Yoshie, have been diligently working ever since introducing numerous electromagnetic tools to their Japanese audience through their books, publications, consulting, and many, many training classes. I was honored to have our software included in their efforts.

This is one of their books. Aki and Yoshie found an insatiable thirst for information on antennas. The need and desire to learn was overwhelming, but the strengths of their students often lay well outside this field. Antenna knowledge was minimal. Aki and Yoshie have over the decades encountered literally thousands of these bright, eager students in Japan and they have quickly brought them up to speed on this often highly technical and critically important topic.

This book is the culminating product of all those decades of training and research. The typical engineering antenna book is highly technical, requires a reasonable knowledge of electromagnetics, and absolutely no fear of equations. That kind of book is excellent for a good engineering graduate student. It is completely unsuitable for most of Aki and Yoshie's audience.

In response to this need, Aki and Yoshie have developed a curriculum that teaches most of the important basic antenna concepts. There are a few equations, but all topics are introduced from a physical, understanding point of view. The goal is to understand what is going on. If our reader then wishes to pursue the intense mathematical treatments in the usual textbooks, she will now find a much deeper understanding. Equations are always easier to understand if you understand the concepts and their physical meaning first, something Clerk Maxwell himself encouraged. If, instead of becoming an antenna expert, our reader is working with antenna experts; he can now communicate with them, on their own terms, in their own language, and actually understand what they are talking about.

Aki and Yoshie are the authors of the original Japanese version of this book. I have contributed some ideas here and there, but it is otherwise entirely their own work. For this edition, Aki and Yoshie performed an initial translation into English. My efforts were only to put some final polish on the translation, adding just a little bit as I felt appropriate.

Previous introductory books by the Kogures have sold tens of thousands of copies in Japan. With the English edition of this new book, you will now see how much of the Japanese technical community is gaining their initial knowledge of antenna design. You can now enjoy this same introduction. As a special treat, you can see antennas from the Japanese point of view. Note the subtle pride and imagine how a Japanese reader would feel reading about, for example, the developments of Prof. Uda and Yagi.

If you are so kind as to enjoy this book, I speak for Aki and Yoshie Kogure as well as for myself by saying, "Arigatou gozaimashita," or "Thank you very much."

James C. Rautio
March 2011

1

The Antennas Around Us

1.1 What Is an Electrical Circuit?

Electric current always and forever flows in complete loops. This basic bit of information is sometimes lost in the details of advanced antenna design, resulting in surprises when an antenna or circuit is turned on. Let's review this basic concept.

1.1.1 Circuit with Two Parallel Lines

Back in elementary school, we used two wires to connect a battery to a light bulb. The electrical appliances we use also have two wires (and possibly a third, but that is for safety). Thus, electricity requires a complete loop, going out one wire, and coming back the other. This is true for both direct current (DC) as with a battery, and for alternating current (AC) as in our electrical appliances. Both DC and AC systems are divided into three parts: the power supply (the battery, or wall outlet), the transmission line (the two parallel wires), and the load (lamp, or electrical appliance); see Figure 1.1. The devices that are sent electricity are typically called loads and include, for example, lamps, resistors, and integrated circuits (ICs).

The two parallel wires might be called a transmission line. They conduct the electricity from the power supply to the load and back, forming a complete loop. These three elements together are called the electric circuit, or simply the circuit.

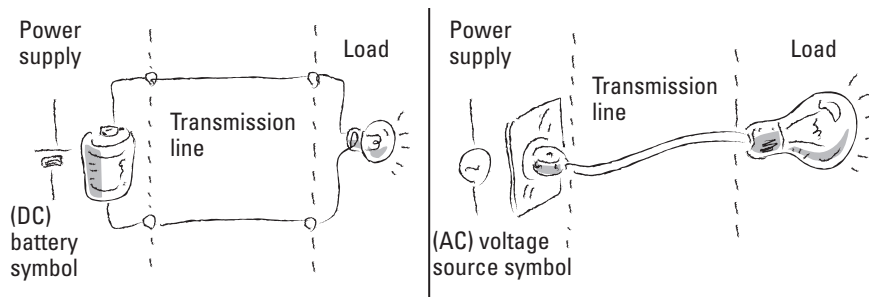


Figure 1.1 Three elements of an electrically powered circuit: power supply, transmission line, and load (lamp).

1.1.2 Role of the Ground Conductor

In Figure 1.2, aluminum foil is used instead of the second wire. We call the entire sheet of foil a ground, sometimes abbreviated GND in diagrams. The term “ground” is similar to “earth,” and in fact sometimes circuits are grounded by connection to the earth. However, the earth is usually a very poor conductor and cannot function as the second, electric current returning wire of a circuit. We must always include our own ground formed from a good conductor, either in the form of a wire (Figure 1.1), or in the form of a ground plane (Figure 1.2).

Every electrical system has multiple complete circuits. If we always actually wire up both wires of every circuit, it gets complicated. There is no reason we can’t use the same return wire for all the circuits. This cuts the total number of wires in half. If this single return wire is a sheet of metal (Figure 1.2), then we call it a ground plane.

1.1.3 Antennas at the Edge of a Substrate

Figure 1.3 shows the inside of a WiMAX transmit and receive module. There are many circuits formed by the thin copper foils (called lines or traces) on the surface of a thin insulating substrate.

The antenna consists of a pair of meander lines in the upper corners with some small surface mount technology (SMT) or surface mount devices (SMDs) chips near the bottom of the antenna. This board is called a printed wiring board (PWB) or a printed circuit board (PCB). Note that the part called an antenna is on the edge of the board, and to the untrained eye, it looks just like the rest of the circuitry.

The two antennas are separated by a ground conductor in the shape of Mt. Fuji. There is a ground conductor on the reverse as well, and it is also called a ground where it electrically extends to the rest of the PCB.

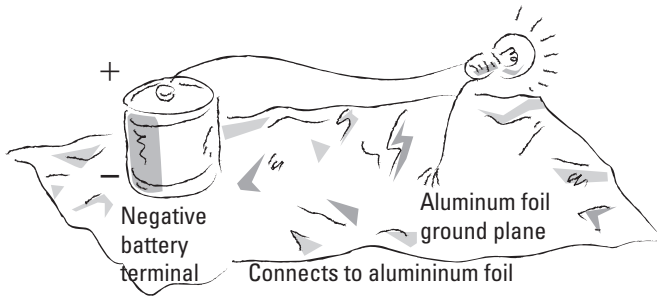


Figure 1.2 Ground conductor made with aluminum foil.

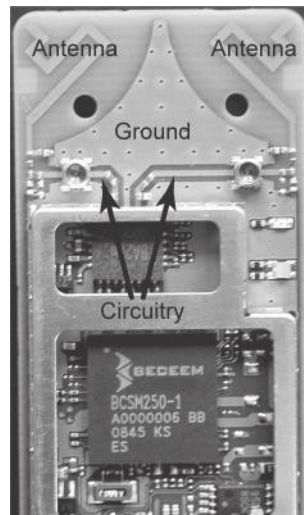


Figure 1.3 Transmit and receive module of a WiMAX system.

There is no ground conductor directly underneath the antenna traces. For good radiation, it is important to put as few extraneous conductors as possible in the vicinity of the antenna. The many small dots on the Mt. Fuji shaped ground plane are vias. They connect the front surface to the back to make one solid ground plane.

1.2 Just Exactly What Is the Antenna?

The antenna performs the function of transmitting and receiving electromagnetic waves. An antenna is usually made from metal pipe, conducting wire, or,

as in this case, metal traces on a PCB. It is usually composed of good conductor so that it can convert a received electromagnetic wave into electric current, which is then delivered to an electric circuit. For transmit, the antenna converts an electric current into an electromagnetic wave.

1.2.1 Television Antennas

When looking at the front or back window of a car, you might see thin lines embedded in the glass. These are the antennas for your radio, television, and other wireless services. They are formed with a film of conductor on or in the glass.

Figure 1.4 is an example of a common amateur radio antenna, a Yagi antenna. This type of antenna is also used for TV and FM. The elements of a Yagi antenna (the “ribs” in Figure 1.4) convert an electromagnetic wave to electric current most efficiently if they are of the correct size for the frequency being received (or transmitted). The same is true for the antenna in your car window, even though the shape and appearance is completely different. In both cases, the antennas are made from conductors that are shaped and sized to efficiently receive and transmit the desired frequencies.

The wiring at the far end of the Yagi antenna is a coaxial cable connecting to one element. All the rods, or ribs, are called elements and the special characteristic of the Yagi antenna is that there is an electrical connection to only one element. The full proper name of the antenna is a Yagi-Uda antenna, but it is frequently just called a Yagi. It is named after its inventors, Dr. Yagi and Dr. Uda, and is well-known worldwide.

The backmost set of metal rods form a special element called a reflector, which catches a portion of the electromagnetic wave that gets past the front part of the antenna and reflects it back to the antenna so it can receive the signal even better. The purpose of all the elements in this antenna is to receive a signal

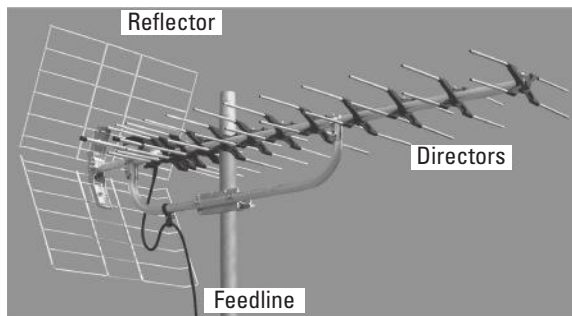


Figure 1.4 Example of a Yagi antenna for FM broadcast reception.

strongly in one direction, but to also reject any interfering signals in other directions. This is a directional antenna and it must be pointed at the station that is to be received.

1.2.2 Antenna in a Radio-Synchronized Clock

Figure 1.5 shows the inside of the radio-synchronized wristwatch (made by Casio Computer Co., Ltd.). The thin stick in the lower side is a coil of very fine enameled wire wound on a ferrite core.

Looking carefully near the ends of the winding, you might see the tiny chip capacitors soldered into the circuit. These parts form an electric circuit called an LC resonator. It is designed to resonate at a specific frequency. “L” stands for an inductor (the coil of wire) and “C” stands for the capacitor (the chip capacitors soldered onto the inductor). The LC circuit can resonate electrically just like a tuning fork can resonate mechanically.

The radio-synchronized clock displays the correct time by receiving an accurate time signal. In Japan, the time signal comes from either of two standard frequency broadcasts. One is transmitted at a power of 50 kW and on a frequency of 40 kHz on a 250 meter tall antenna on Mt. Ohtakadoya-yama in Fukushima Prefecture. The other signal is on a frequency of 60 kHz using a 200 meter high antenna at the border between Saga Prefecture and Fukuoka Prefecture, which belongs to the National Institute of Information and Communications Technology (NICT).

Each side of the WiMAX antenna in Figure 1.3 has a length of about one-quarter of a wavelength. The size of the element of the Yagi antenna in Figure 1.4 is about one-half of a wavelength. If we know the frequency of the electromagnetic wave, we can calculate the wavelength. Then we just make the antenna the appropriate fraction of a wavelength long and it works well. However

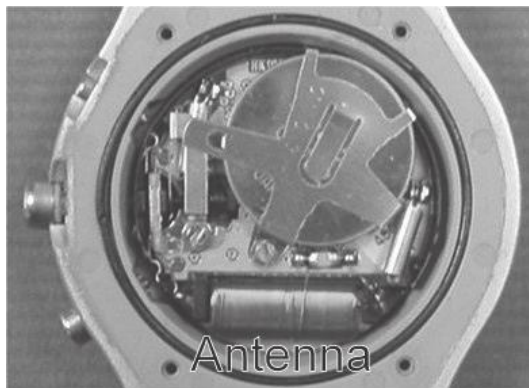


Figure 1.5 Antenna for a radio-synchronized wristwatch.

the wavelength of a 40-kHz electromagnetic wave is about 7.5 km and we cannot put an antenna several kilometers long inside a wristwatch. Perhaps if we unwrap the coil of the inductor in Figure 1.5, could it possibly be that long?

1.2.3 Is the Coil of a Radio-Synchronized Clock an Antenna?

Figure 1.6 shows an experiment first conducted by the physicist Hans Christian Oersted of Denmark (1777–1851). We see that the magnetic needle moves when the current flows in the electric wire. Thus, he discovered that electric current generates the magnetism. He did not understand why, but his results triggered the subsequent development of electromagnetics.

Based on Oersted's work, the physicist André-Marie Ampere of France (1775–1836) discovered that the direction that the magnetic needle moves depends on the direction that the current flows. Ampere's right-hand screw rule (right-hand rule), shown in Figure 1.7, is that if you point the thumb of your right hand in the direction of the current, the resulting magnetic force is in the direction that your fingers curl.

The miniature antenna of the radio-synchronized wristwatch in Figure 1.5 is a dedicated receiving antenna. It can receive the time signal because the

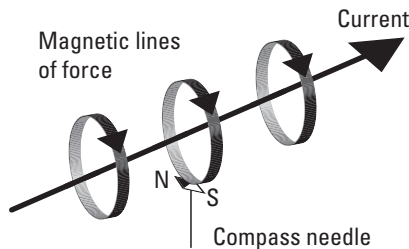


Figure 1.6 Oersted's experiment.

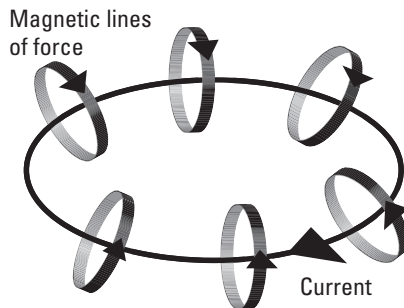


Figure 1.7 Ampere's right-handed screw rule.

magnetic field of the electromagnetic radiation transmitted in free space concentrates in the coil with the help of the ferrite bar (magnetic material), the core around which the fine wire is wound.

The ferrite core helps to generate electricity from the magnetic portion of the received electromagnetic wave using the principle of Faraday's law of magnetic induction. Michael Faraday (1791–1867) discovered that a changing amount of magnetism generates electricity in a coil of wire. To improve reception, designers added capacitors to the coil of wire in the wrist watch to make it resonant at precisely the desired frequency.

Because the coil used for this purpose might be called an antenna in the industry, it is also included as an antenna in this book.

1.3 Fundamental Form of Antennas

Antennas come in a very wide range of physical realizations. However, most antennas of interest for this work can be grouped into one of the following categories.

1.3.1 The Yagi Antenna

The Yagi antenna is sometimes used for optimal reception of specific television channels, with the feed line connected only to one element. (This is different from a broadband log-periodic antenna, which can receive a large number of channels.) A Yagi can also be used to receive FM broadcasting. Because the frequency is lower (and thus the wavelength is longer), the FM antenna is larger than the television antenna. A two-element Yagi for 7 MHz is shown in Figure 1.8. There are several other smaller Yagi antennas (for higher frequencies) above it.

Figure 1.9 shows an antenna that uses four metallic rods. Each of the rods is one-quarter of a wavelength long. There are two one-quarter wavelength rods for each of two elements, A and B. Thus, each element is one-half of a wavelength long. The centers of each element are both connected to the transmitter by the parallel feed lines. As the parallel line spacing is narrow, they do not radiate. Only the two elements, A and B, radiate. However, this is not a Yagi antenna. Let's see how it works.

To understand how an antenna works better in some directions and worse in others, we must visualize electromagnetic waves propagating through space. If you have not done this before, just imagine sound waves or water waves. When you have two waves, depending on their timing and direction, they can add in phase, or they can cancel.

The element spacing for this antenna is $\frac{1}{4} \lambda$ (λ = wavelength). One-quarter wavelength is 90° of a sine wave. Thus, element A starts radiating first, then,

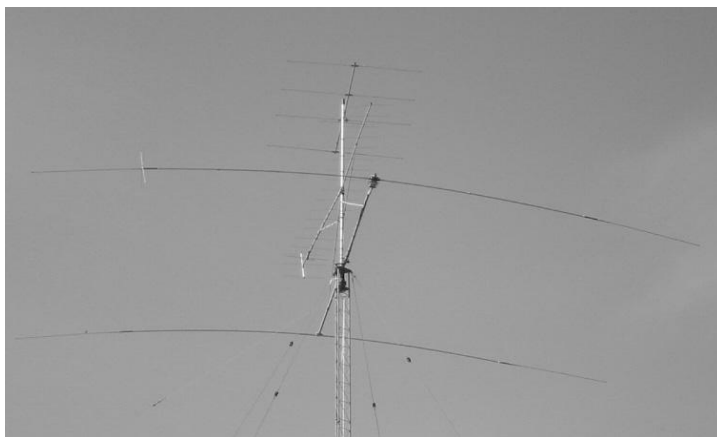


Figure 1.8 A two-element Yagi for 7 MHz. (Courtesy of Ron McClain, W2VO.)

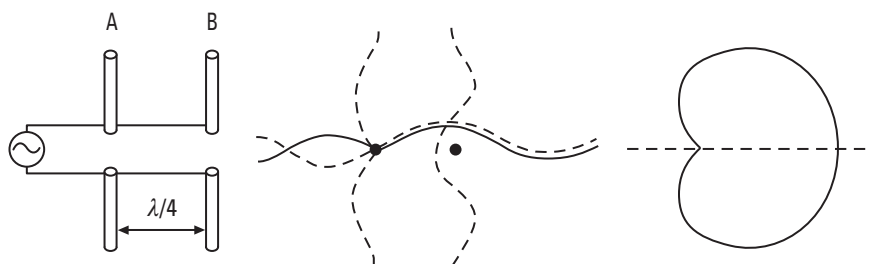


Figure 1.9 Antenna of two elements (cardioid pattern).

element B starts radiating later. Pretend that we are a receiver, standing far to the right. We will receive two electromagnetic waves (Figure 1.9(b)), one from element A, and the second from element B. Element A radiates first, but it must travel an extra $\frac{1}{4}\lambda$ to get to you. The wave from B starts out later, but it is $\frac{1}{4}\lambda$ closer to us. Thus we receive both waves at the same time and they add constructively. We get a strong signal.

Now pretend we are a receiver standing far to the left. The signal from element A is radiated first and gets to us first. The signal from element B radiates later (delayed by the $\frac{1}{4}\lambda$ transmission line) and must travel an extra $\frac{1}{4}\lambda$ to get to us. Thus it is delayed by $\frac{1}{2}\lambda$. It is perfectly out of phase with the wave from element A and the two waves cancel. We receive nothing.

When we mathematically perform the above analysis for all possible directions, the radiation pattern of the entire antenna is a heart-shaped (cardioid pattern); see Figure 1.9(c).

Figure 1.10 shows the radiation pattern when the phase difference between A and B is adjusted to 120° . Comparing with the cardioid pattern in Figure 1.9(c), we still have strong radiation to the right, but the directivity is higher. In other words, signals off to the side will be weaker. Because the electromagnetic field simulator used in this book can analyze antennas for a variety of spacing and the phase differences, we can use it to design an antenna for the desired characteristics.

Dr. Yagi and Dr. Uda discovered that directivity is obtained even when the feeder wire between A and B is removed, as shown in Figure 1.11. This is because whenever current flows in element A, some electric current is then also induced in element B (by Faraday's law, mentioned above).

In fact, Profs. Uda and Yagi also discovered that directivities increase even more when they use more elements, but only if they changed each element length slightly (Figure 1.12).

1.3.2 Electromagnetic Simulation for Antennas

The Yagi antenna is designed to obtain a maximum value of gain by adjusting the size and the position of elements carefully. Now let's investigate the two-element Yagi antenna with electromagnetic field simulator Sonnet Lite™. Sonnet Lite is available online at www.sonnetsoftware.com, or can be installed

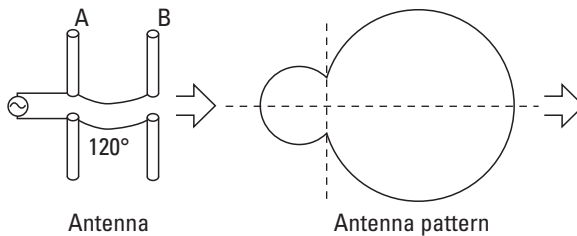


Figure 1.10 Antenna of two elements with a phase difference between A and B of 120° .

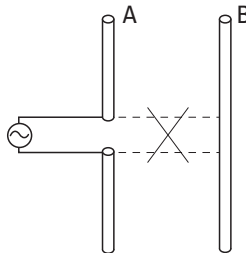


Figure 1.11 Feeder wire between A and B is removed.

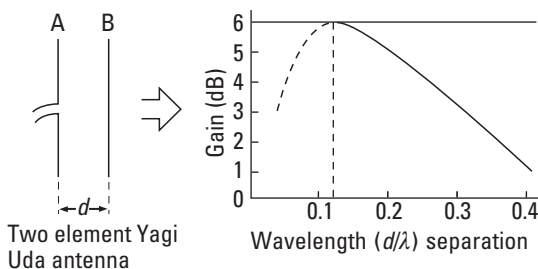


Figure 1.12 Directive gain of a two-element Yagi antenna. Maximum gain is obtained spacing at 0.11λ .

from the DVD provided with this book (see the Appendix for details). First, we draw a simple antenna, shown in Figure 1.13.

Input the Element

To start Sonnet Lite, click the Edit Project button of Task Bar and select New Geometry; an initial screen similar to that in Figure 1.14 is displayed. This rectangle represents the surface of the substrate viewed from above. This is where we will draw our antenna.

The EM analysis will mesh our circuit. The fine grid of points on the substrate surface shows the minimum mesh size (user definable). This can be

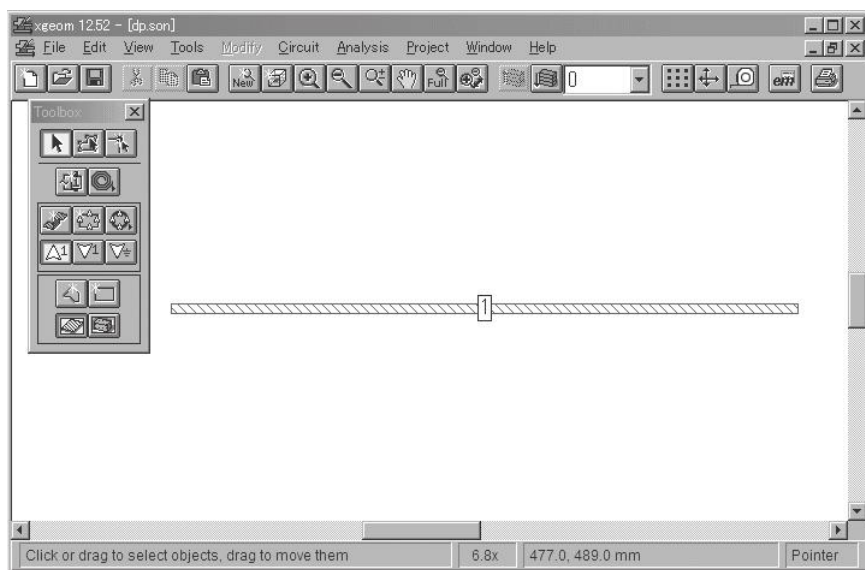


Figure 1.13 The antenna element we will draw and analyze, file name: dp.son.

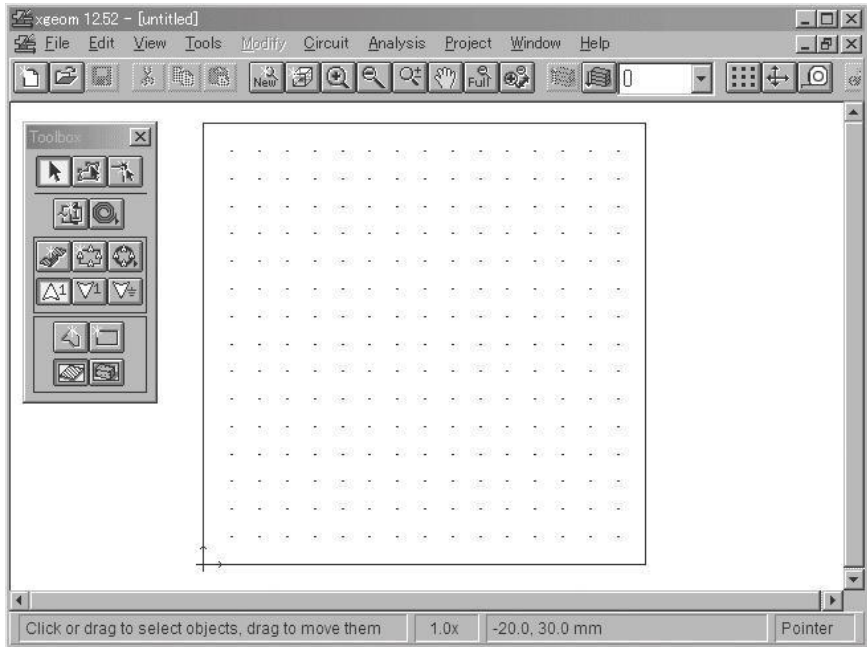


Figure 1.14 The initial screen looking down at the top surface of the substrate.

viewed as a snap-grid. The rectangle formed by four closely spaced points is the minimum mesh size and is called a cell.

After selecting Circuit->Units..., the dialog box in Figure 1.15 is displayed. Set mm as length units and MHz as frequency units.

Next, select Circuit->Box... and the dialog box in Figure 1.16 appears. Set the size of both the x and y direction minimum mesh to 1 mm. Then set the size of the substrate, and when you input 1,024 mm in x and y , respectively, number of cells (Num. Cells) is automatically updated.

Next, set both the Top Metal and Bottom Metal, upper right corner of the dialog box, to Free Space.

These settings allow the analysis to include electromagnetic wave radiation.

To draw the antenna element, first click the button in the lower right of Toolbox shown in Figure 1.17. Next, put the mouse cursor on the corner of a cell, drag while holding down the left button, and release the button when your rectangle is the correct size, which is 4 mm (four cells) wide and 126 mm long, about at the center of the substrate. The actual size is displayed as you are drawing it, in the lower right corner of the Sonnet display. We now see a rectangle with red hatched lines (lossless conductor).

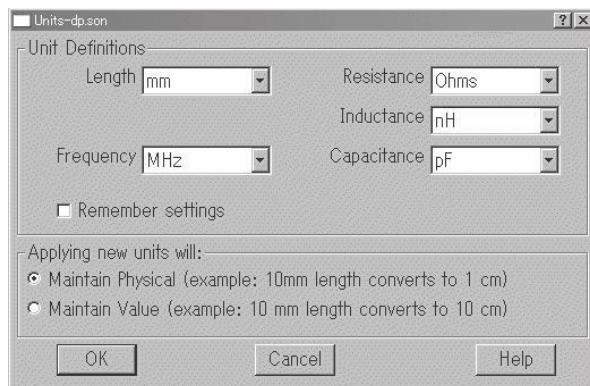


Figure 1.15 After selecting Circuit->Units..., this dialog box is displayed.

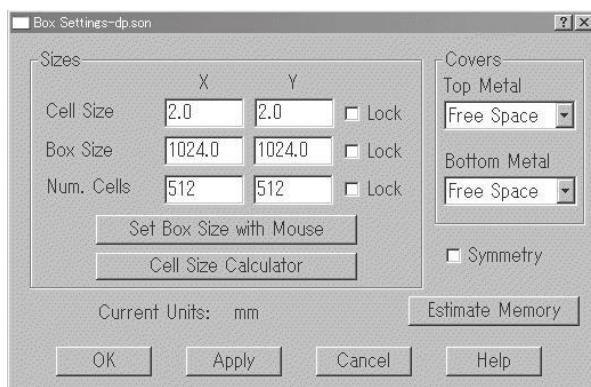


Figure 1.16 Dialog box displayed after selecting Circuit->Box... to set the cell size and the box size.

Because this only one-half of the element, draw another element in the same way, and arrange it so that the edges of two elements touch at the center, as shown in Figure 1.18.

Another method of drawing a rectangle is to input the size directly and click the OK button in the dialog box shown in Figure 1.19, which appears after selecting Tools->Add Metalization->Rectangle.

If you don't get the rectangles in quite the right place at first, just left-click on one and drag it to where you want. To exactly center the entire antenna in the substrate, select the entire antenna and then choose Modify->Center->Both.



Figure 1.17 The Toolbox.

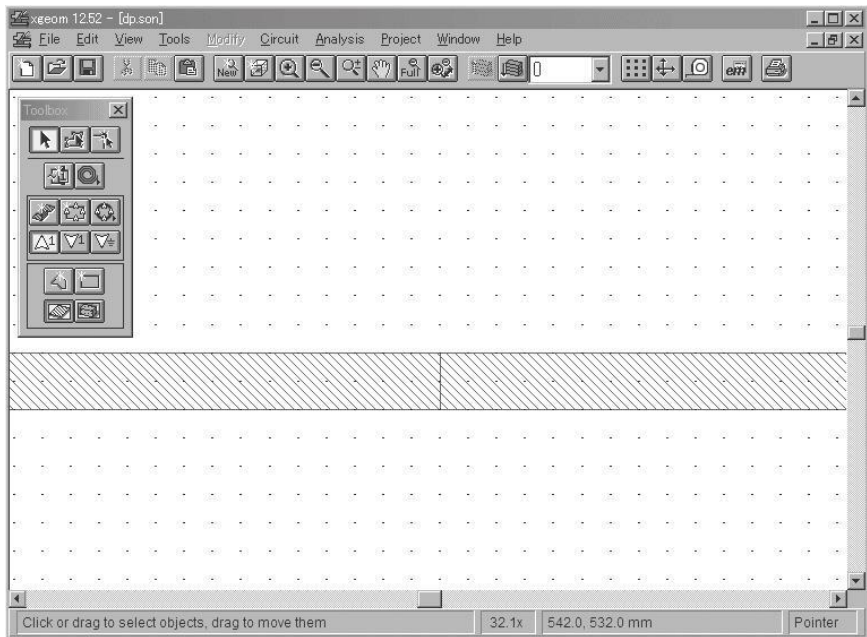


Figure 1.18 Draw two rectangles so that they meet at the center of the box.

Other Settings

Next, click the button on the second row on the left of Toolbox, as shown in Figure 1.20. Then click where the two rectangles of Figure 1.18 meet to create a port. A port is where we would connect, say, a coax connector, to attach

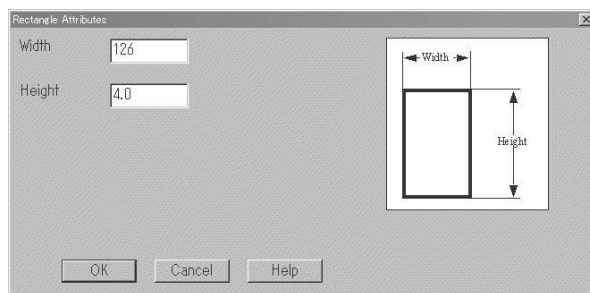


Figure 1.19 Dialog box to draw rectangle that is 125 mm wide and 4 mm long.



Figure 1.20 The Add Port button.

a transmission line to feed the antenna. Imagine that there is a very tiny gap between the two rectangles, right where we placed the port. The port allows the EM analysis to put a voltage across that tiny gap and see what happens. When we connect a coax cable, the center conductor is connected to one side of the gap, and the shield conductor is connected to the other side.

The default metal for the antenna element is lossless. For a more realistic situation, we now change the metal to copper. Select Circuit->Metal Types... and then click the Add... button on the dialog box that appears. Next click Select metal from library... in Figure 1.21 and click Global Library. A list similar to that in Figure 1.22 is displayed. Here select Copper and click the OK button. Next, input, 0.03 mm in the Thickness box, shown in Figure 1.23, and set the name to Copper.

After double-clicking on the antenna element, the dialog box shown in Figure 1.24 displays. Change Lossless to Copper in the Metal item. Be sure

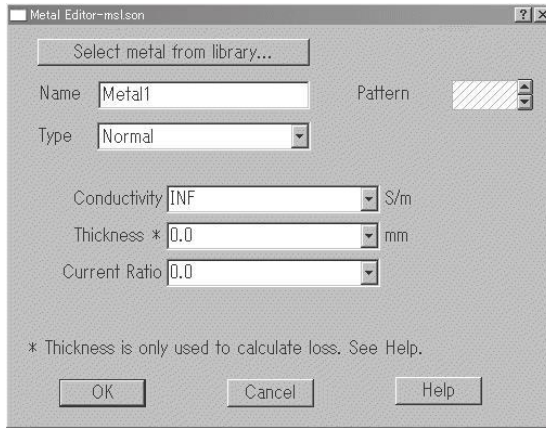


Figure 1.21 Click the Select metal from library... button.

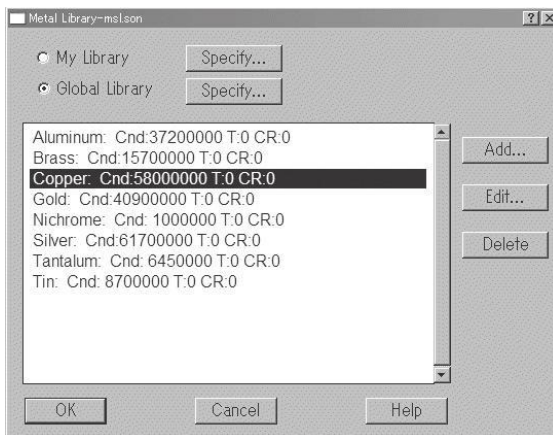


Figure 1.22 Select Copper and click the OK button.

to change both rectangles. If you select them both at the same time, you can change both at the same time.

Next, we set the height of the space above and below the substrate (z direction). The sizes of the x and y directions are already set in the box of Figure 1.16. Figure 1.25 is an input screen of the thickness of the layer in which the space being analyzed is viewed from the side. It is set in the dialog box displayed after selecting Circuit->Dielectric Layers.....

We have been drawing the wiring pattern on the surface of a dielectric substrate layer. In Sonnet, these surfaces are called Levels and each level has a

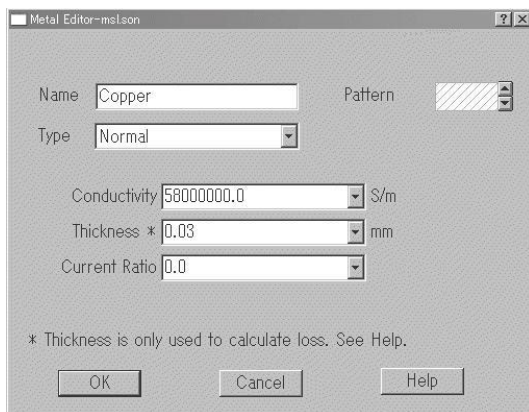


Figure 1.23 Input Thickness as 0.03 mm.

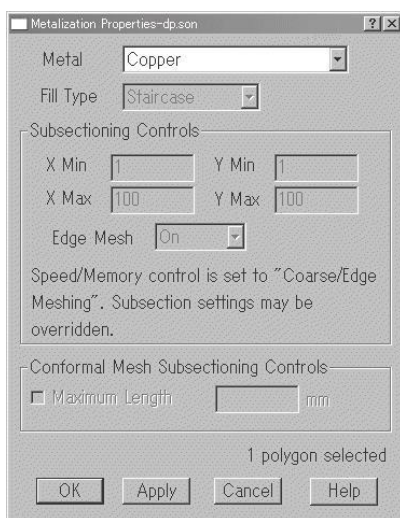


Figure 1.24 Change Lossless to Copper in Metal item.

number. The antenna element in Figure 1.18 is on Level 0. Because the layers above and below Level 0 are set to 200 mm as shown in Figure 1.25, the height of space being analyzed is 400 mm.

Here, we set only the thickness of dielectric layers. The parameter ϵ_{rel} means relative permittivity (ϵ_{rel}); the default of 1.0 represents air. Since there are only two dielectric layers, and they are both air, this antenna is floating in air. A real antenna will be fabricated on an actual physical substrate. In this case,

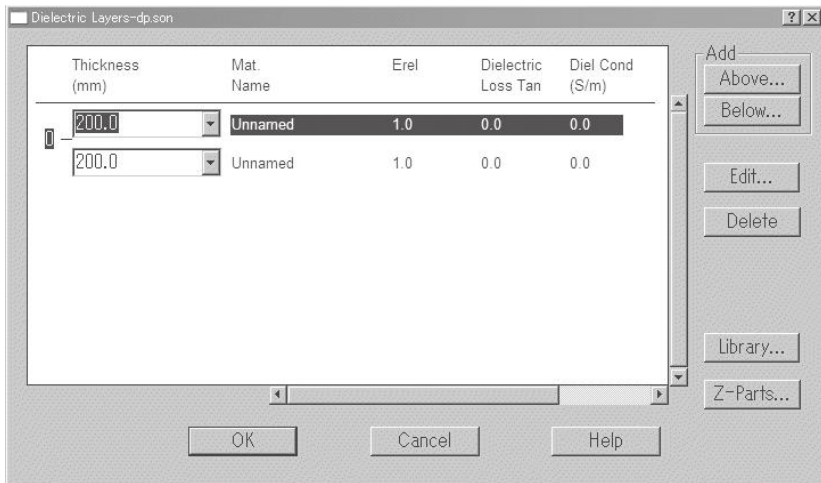


Figure 1.25 Set the height of space above and below the substrate (z direction) by the setting the thicknesses of the dielectric layers.

the designer obtains the dielectric constant of the substrate from the substrate vendor, and then adds a third dielectric layer with that dielectric constant and an appropriate thickness. We are not doing this in order to simplify this first example.

This model represents an antenna appropriate for a single digital terrestrial television. The size we have selected means that it resonates at around 600 MHz, where the wavelength is 50 cm long. It is calculated when dividing 3×10^8 m/s by 600×10^6 Hz where the numerator is a velocity of electromagnetic wave (velocity of light), so the height of our analyzed space is about one wavelength. A guideline for Sonnet suggests that all edges (top, bottom, and sides) of the space being analyzed should be around one-quarter to one-half a wavelength from the antenna element.

A 3-D view is displayed by selecting View->View 3D (see Figure 1.26).

Run a Simulation

After inputting all the data, save the model by selecting File->Save (users of the Asian versions of Windows should input the name “dp” in single-byte characters, not double-byte.) Sonnet then makes the file dp.son with the extension .son and a folder named sondata.

In the dialog box (Figure 1.27) displayed after selecting Analysis->Setup..., input 500 MHz for the Start frequency and 700 MHz for the Stop frequency. The default Adaptive Band Sweep (ABS) needs only the range of frequency for analysis and we trust Sonnet to decide the frequency step and to correctly and accurately interpolate results from EM analysis at just a few frequencies.

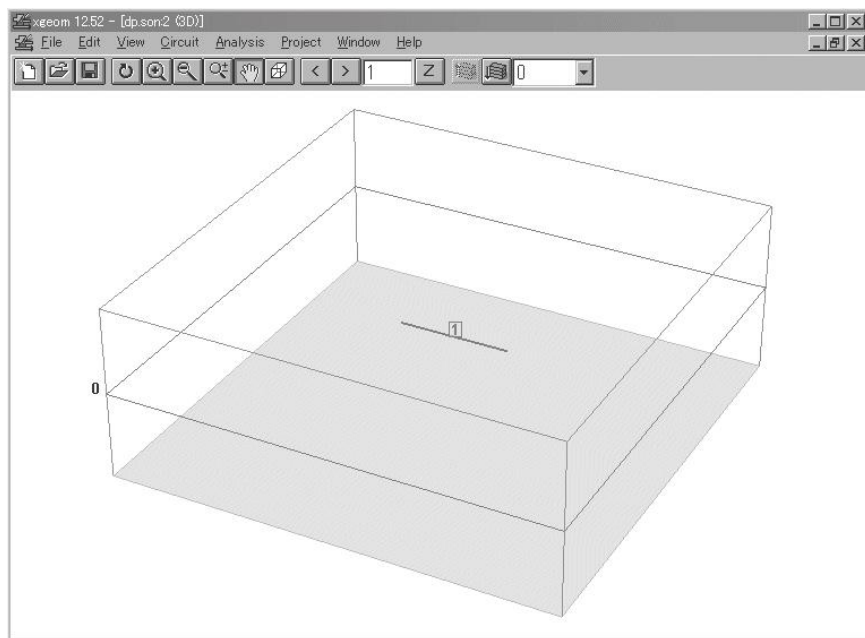


Figure 1.26 A 3-D view of this antenna model.

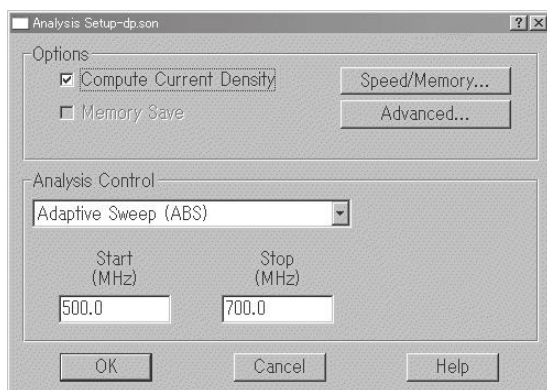


Figure 1.27 Set the range of frequencies for simulation.

Alternatively, we can also set a Linear Frequency Sweep that runs a simulation using user-specified start, stop, and step frequencies.

Furthermore, if you check Compute Current Density in Options on the left top, you can view the current density distribution on the surface of the metal element at a later time.

Figure 1.28 shows the window that appears after selecting Project->Analyze, which scrolls up during the simulation. When the progress bar reaches 100%, selecting Project->View Response->New Graph displays a plot similar to that in Figure 1.29.

This is a plot of the expected reflection coefficient, or magnitude (dB) of S_{11} .

S-parameters, an abbreviation of scattering parameters, are often used with high-frequency circuits. When there is only one port, as with this antenna, there is only one S-parameter, S_{11} . The plot in Figure 1.29 shows the voltage of the wave reflected from port 1 (converted to dB), when a 1-volt amplitude wave is incident on port 1. Or, more generally, if the incident wave has a voltage magnitude of a_1 , and the reflected wave has a magnitude of b_1 , then S_{11} is the magnitude of the ratio of b_1 over a_1 . The value on the graph is obtained by converting that ratio to dB. We call it return loss.

$$\text{return loss} = 20 \log_{10} \left| \frac{b_1}{a_1} \right| = 20 \log_{10} |S_{11}| [\text{dB}]$$

Now, for some examples. If the entire incident wave is completely reflected, then the S_{11} magnitude is 1.0. When we convert to dB, the return loss is 0 dB. Of course, no power is radiated by the antenna because all of the input power is reflected.

Let's say that the voltage of the input wave is 1 volt and the reflected wave is only 0.1 volt. Now, the S_{11} magnitude is 0.1, which is the same as -20 dB.

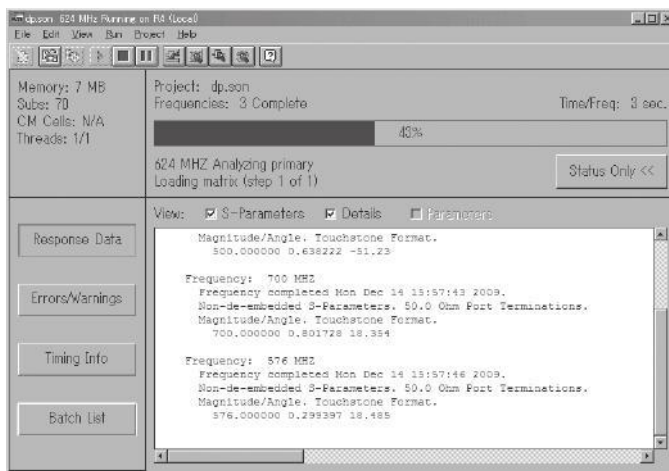


Figure 1.28 The process monitor window is displayed during simulation. The results scroll up the screen.

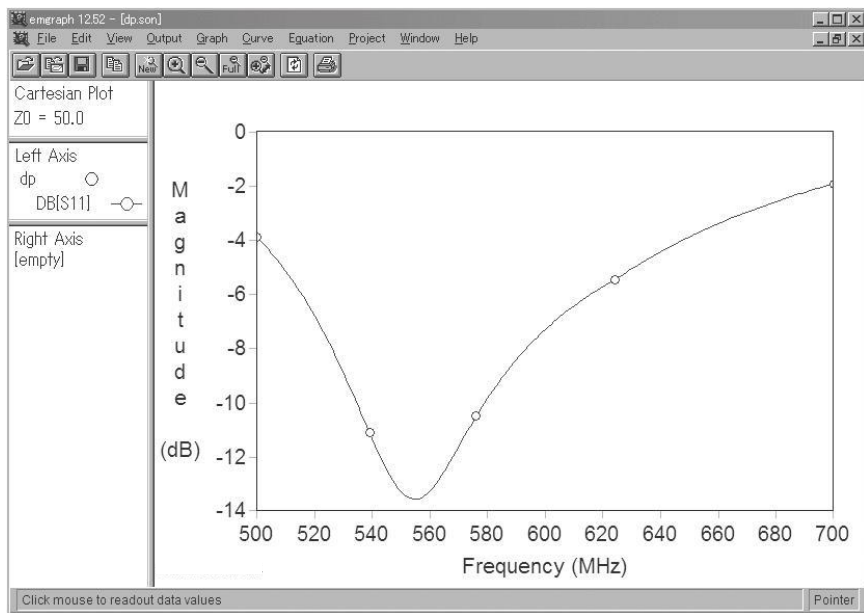


Figure 1.29 Simulation result: a plot of S_{11} (reflection coefficient).

With a return loss of 20 dB, very little power is reflected from the antenna. That means that most of the incident power is either radiated or absorbed in antenna losses.

Sharp-eyed readers will note a sign problem in the above example. Strictly speaking the return loss is 20 dB, but the equation produces -20 dB. You will find that this sign mistake is common in practice. Just keep in mind that all passive antennas must reflect less power than what is incident.

If you are not comfortable with dB, change Data Format from Magnitude (dB) to Magnitude in the dialog box (Figure 1.30), which will appear after double-clicking DB[S11] shown inside the frame on the left. The vertical axis changes to that of Figure 1.31.

What Is the Input Impedance of the Antenna?

The characteristic impedance of a transmission line, like coaxial cable, is the ratio of voltage to current for a wave traveling down the cable. It is determined by the dimensions of the cable and the electrical properties of the material between the conductors. It is important because a traveling wave does not like the impedance to change. If the impedance changes, then a second, reflected, wave is launched. Reflected waves go back to the transmitter, they do not get radiated.

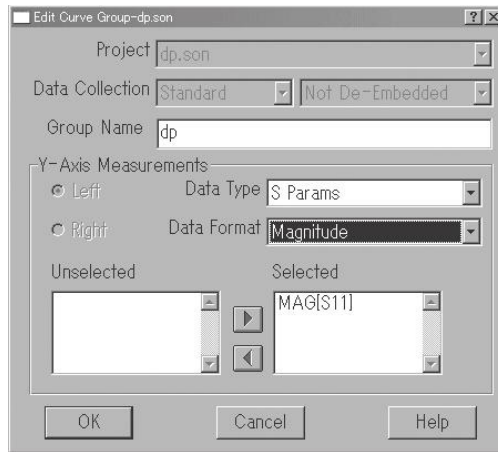


Figure 1.30 Change Data Format from Magnitude (dB) to Magnitude.

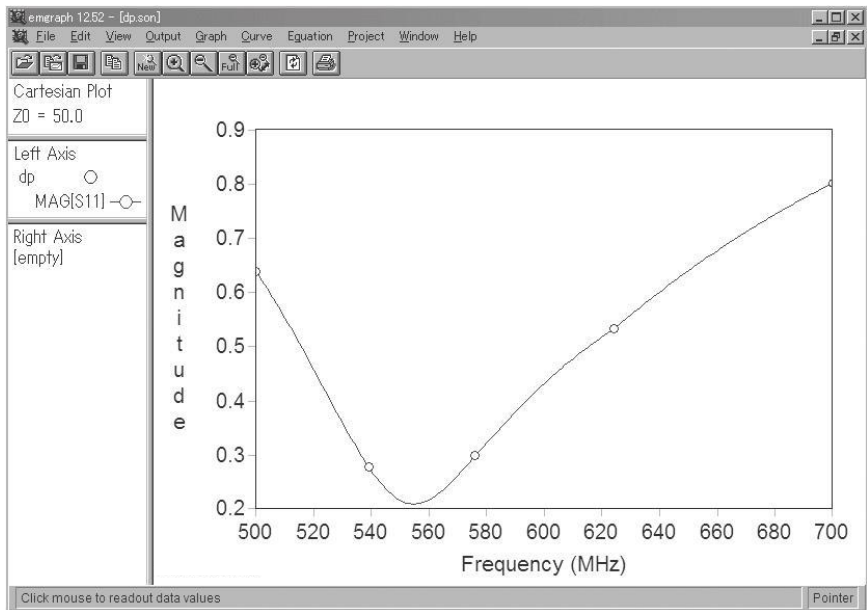


Figure 1.31 The vertical axis is now straight magnitude, not converted to dB. This is the reflected wave voltage when the incident wave has an amplitude of 1V at the input port.

The antenna also has an impedance. The antenna input impedance is the ratio of voltage to current as measured at the antenna input terminals. Ideally,

we want to have all the power traveling along the transmission line to go into the antenna. We do not want any power reflected. To do this, we typically need to have the antenna impedance and the transmission line characteristic impedance to be as close to the same as possible.

The characteristic impedance (Z_0) of coaxial cable used for TV antennas is usually 75 ohms (Ω). Electromagnetic field simulators in general output S-parameters assuming that a 50 Ω transmission line is connected to the port. This is described as S-parameters normalized to 50 Ω , as in Figures 1.29 and 1.31. If we are actually going to connect a 75 Ω transmission line, we need to view S-parameters normalized to 75 Ω (Figure 1.32).

To renormalize, select Graph->Terminations... and enter 75 in the resistance box. After clicking OK, the plot shown in Figure 1.33 is displayed. This now corresponds to the amplitude of the reflected wave when you connect the transmitter to the antenna using 75 Ω cable.

In order to get zero reflected wave as mentioned above, the antenna input impedance must be equal to the transmission line characteristic impedance. So, just what is the antenna input impedance? In the dialog box displayed after double-clicking DB[S11] in the frame to the left of the plot, set Data Type to Zin and Data Format to Real and click OK. The graph for the real part of impedance (R: resistance) appears.

Next, in the frame on the left, double click on the Right Axis label. Set the Data Type to Zin and Data Format to Imaginary and click OK. Now, the imaginary part of impedance (X: reactance) is also displayed. The final plot is shown in Figure 1.34.

In Figure 1.33, the least reflection is at 560 MHz. With minimum reflection at that frequency, we are likely close to maximum radiated power (with some power being absorbed by losses in the antenna). Next, looking at the

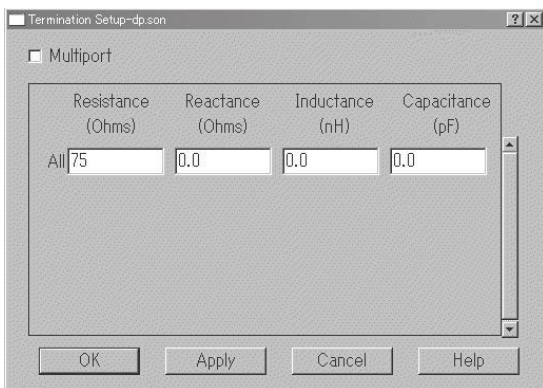


Figure 1.32 Changing the normalizing resistance to 75 Ω .

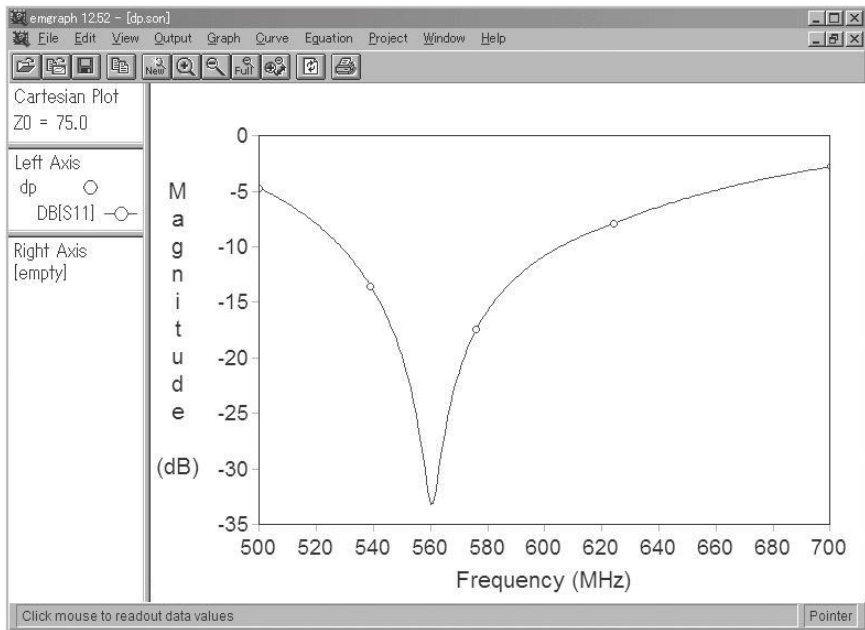


Figure 1.33 The reflection coefficient plot normalized to 75Ω .

impedance in Figure 1.34, the real part is 78Ω and the imaginary part is -1.9Ω . This is usually written as $78-j1.9\Omega$. The imaginary part, the reactance, X , is almost zero and the resistance, R , is almost equal to the transmission line characteristic impedance of 75Ω . Thus, the reflected wave is minimized when the input impedance of the antenna is equal to the characteristic impedance of the feedline and the antenna is said to be “matched,” an important goal in antenna design.

Electric Current on the Antenna

Because we checked the Compute Current Density box in the dialog box of Figure 1.27, we can now view the current density distribution on the surface of the metal antenna element. When clicking Project->View Current in the Sonnet Lite window, the window shown in Figure 1.35 is displayed.

When we view the display at 560 MHz, where we have the best matching, we see the current is strongest in the center and at both ends goes to zero, and the antenna element is one-half a wavelength long.

Simulation of a Yagi Antenna

We provided a detailed description above because this is our first simulation. We next simulate a Yagi antenna. Figure 1.36 shows Sonnet with an antenna

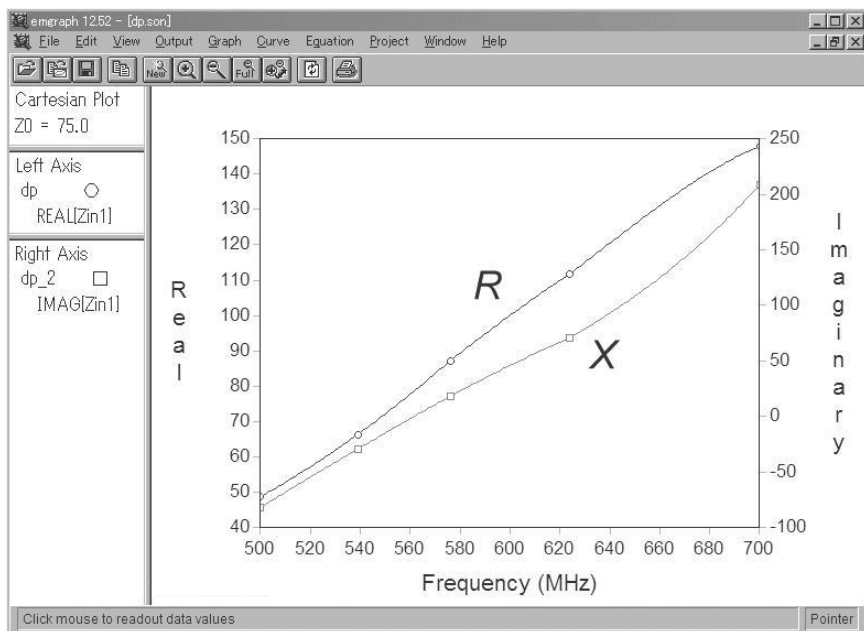


Figure 1.34 Real and imaginary parts of the input impedance of the antenna.

after drawing another element. In Figure 1.12, the directivity is strongest with a spacing of 0.11 times wavelength, and the maximum gain is obtained. So we draw the second element with a spacing of 5.5 cm that is 0.11 times wavelength of 50 cm at 600 MHz. The second element is also called a parasitic element. Ideally, it should be slightly shorter than the driven element, in which case it is also called a director, because it tends to direct power from the driven element.

Figure 1.37 is the result of simulation after selecting Project->Analyze, which shows the S_{11} and the reflection increased. Then, after switching to the graph to view input impedance, you can see that R is now down to about 10Ω around 560 MHz (not shown here). The circuitry required to match a 10Ω antenna input impedance to a transmission line is called a matching network and is a topic of more advanced antenna design. For an example, see Section 8.3.5.

Surface Current of Elements

Display the surface current distribution by selecting Project->View Current and set Animation Type to Time by selecting Animation->Settings... Next, set it to sweep time by selecting Animation->Animate View, then view when the current on the parasitic element is strong, as shown in Figure 1.38.

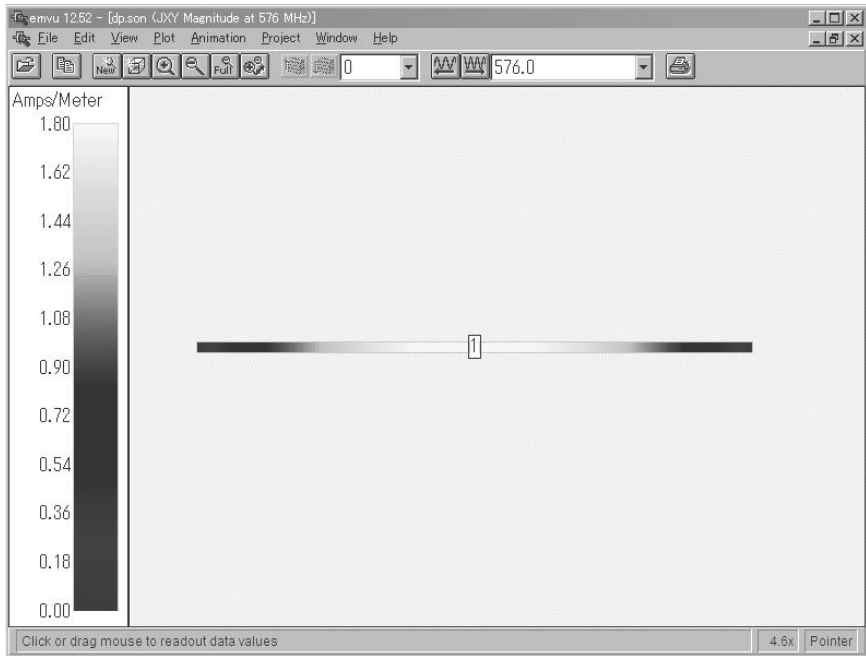


Figure 1.35 Current density distribution on the surface of the metal element. Current at the center is strongest and at both ends go to zero.

Was a Strong Directivity Obtained?

Figure 1.39 is the radiation pattern at 541MHz, which shows the feature of the Yagi antenna (the professional version of Sonnet is required to view this).

We do not obtain such a nice directivity at other frequencies. In fact at some frequencies it actually radiates in the opposite direction (for details, refer to Section 3.3).

1.3.3 Fundamental Forms of Antennas

The one-element antenna in the previous section is called a dipole antenna. It is about $\frac{1}{2}\lambda$ long at its operating frequency of 560 MHz. However, if we calculate the frequency for which the antenna is exactly $\frac{1}{2}\lambda$ long, it should be resonant at 600 MHz. All EM analyses have EM analysis errors. Is this difference due to an EM analysis error? It turns out that this is a real problem and is not due to an EM analysis error. Instead, it is necessary to make the dipole slightly shorter than one-half a wavelength, as we explain in detail in Chapter 3.

The other type of antenna we have looked at is a loop of wire (many loops in the case of Figure 1.5). Another loop antenna is the radio frequency

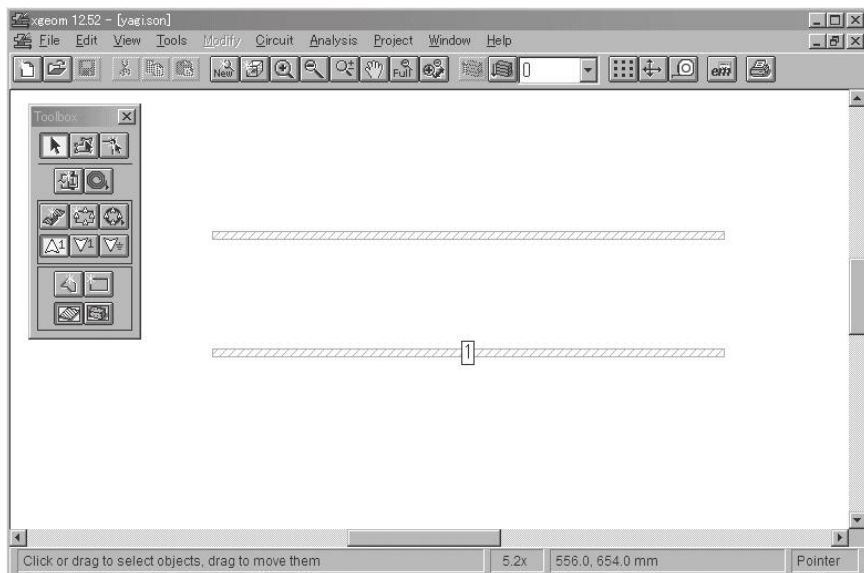


Figure 1.36 Draw the second element and we have a complete Yagi antenna. The spacing is 55 mm.

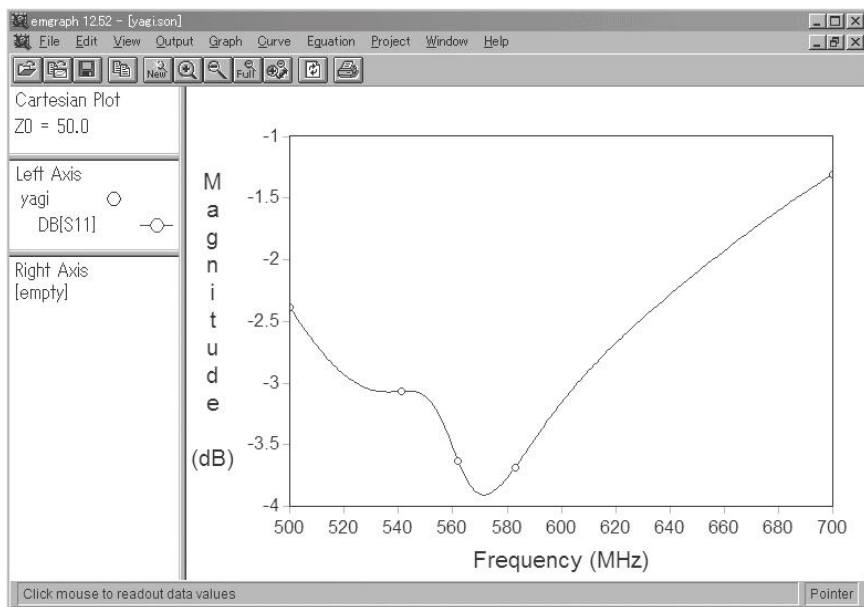


Figure 1.37 S_{11} of the Yagi antenna.

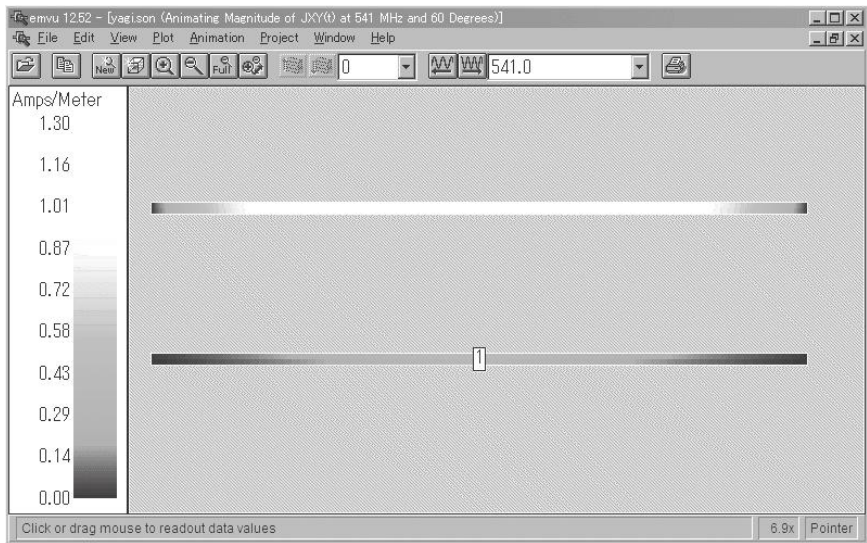


Figure 1.38 Current density distribution in the Yagi antenna.

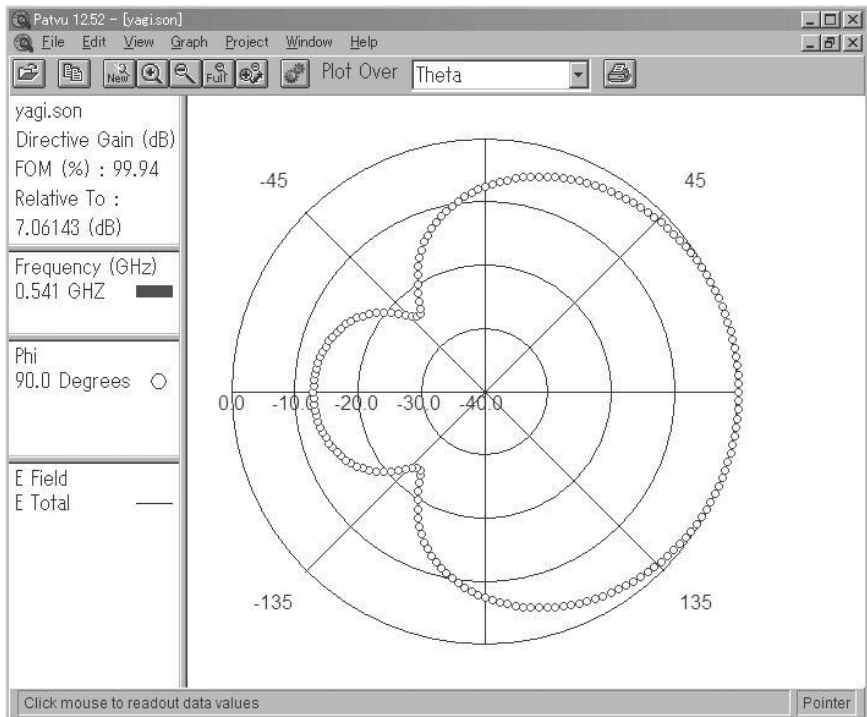


Figure 1.39 Radiation pattern of the Yagi antenna.

identification (RFID) tag, where a silicon integrated circuit (IC) chip is connected to a tiny multiloop antenna inside a credit-card-sized tag. It now sees widespread use in employee ID cards, railway tickets, and in credit cards (Figure 1.40). The cards often use 13.56 MHz. It is impossible to use a $\frac{1}{2} \lambda$ dipole on a credit card because it would be about 10m long. Thus, a tiny loop antenna is used.

All antennas we have described above are classified loosely into two groups, dipole and microloop (Figure 1.41). Because we usually view the dipole antenna as being sensitive to the electric field, we conveniently call it an electric field detection antenna. Strictly speaking, the microloop is not an antenna because it is typically much smaller than a wavelength, and thus cannot radiate very well at all due to its very low input resistance and high input reactance, which forces a matching network to be very narrow band and very lossy. However, over shorter ranges, it can behave like an antenna and is often referred to as such. It is included here because it is so important in today's applications. Because a microloop responds to changes in the magnetic field, it is also called a magnetic field detection antenna (magnetic field antenna).

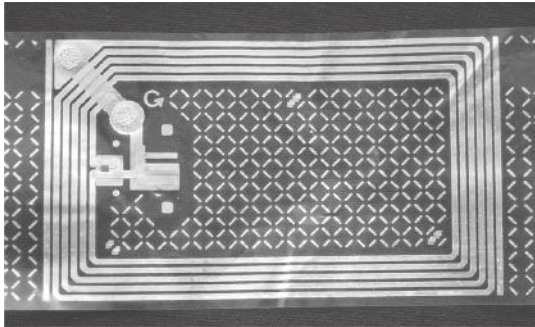


Figure 1.40 An example of RFID tag, with a 5 turn coil.

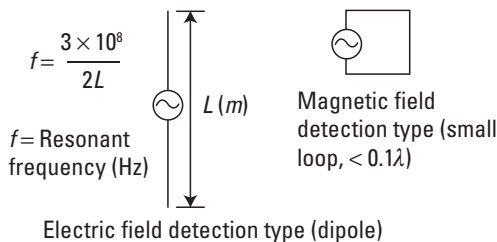


Figure 1.41 Two groups of antennas, the dipole antenna and the microloop antenna.

1.4 What Are Near and Far Fields?

The antenna radiates an electromagnetic field. Near the antenna, the fields are very complicated with energy surging in many different directions. Far from the antenna, the fields are simple, a single traveling wave originating from the antenna. It is important to understand the difference when designing antennas.

1.4.1 Boundary Between Near Field and Far Field

Figure 1.42 shows the electric field strength around a dipole antenna and it also displays the distribution of electric energy at a certain time. This figure is the result of the calculation using the electromagnetic field simulator XFDTDTM by Remcom (USA) that solves Maxwell's equations (see Chapter 2) and displays the electric field and the magnetic field.

Up to about one wavelength around an antenna is called the near field. Electromagnetic energy is very concentrated in this area, and we see the especially intense electric field surrounding the $\frac{1}{2}\lambda$ long resonant antenna.

On the other hand, more than about 1λ from the antenna, circular peaks and valleys appear. Electromagnetic radiation is a wave, and we can see the periodic electric field nulls, and they appear every $\frac{1}{2}\lambda$. Because this electromagnetic wave extends from here on out into the surrounding space (getting weaker as it goes further out), we call it the far field. The far field is a radiation field, and

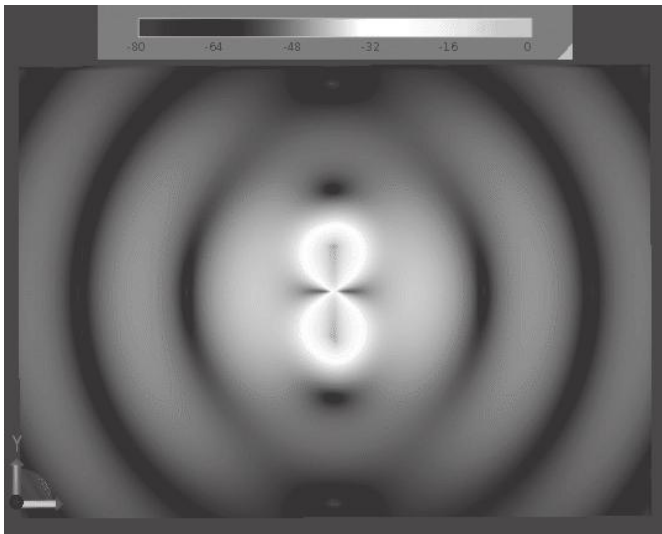


Figure 1.42 Transition between the near and far fields of an antenna. This plot shows only the electric field strength. There is also an accompanying magnetic field. In addition, both electric and magnetic fields have a direction, also called a polarization.

that is what we use to communicate over long distances. In Figure 1.42, you can see that there is no clear boundary between the two. However, it is clear that there is incredible beauty hidden in the universe of Maxwell's equations, as we describe in Chapter 2.

2

Antennas and Radio Waves

2.1 Great Inventions

The field of radio communication started in the 19th century based on a series of amazing inventions. In this section, we briefly go over how we came to be where we are today.

2.1.1 The Experiments of Hertz

The physicist Heinrich Hertz of Germany (1857–1894) (Figure 2.1), using the apparatus of Figure 2.2, demonstrated the existence of the electromagnetic radiation that Maxwell had predicted. He connected a wire from both ends of the induction coil and tied it to small metallic balls separated by a small gap. Then he put large metal plates or spheres on rods extending outward.

2.1.2 Hertz's Receiving Equipment

Figure 2.2 is an apparatus that corresponds to the transmitter and transmission antenna. But how did he observe the electromagnetic wave that had been predicted to exist?

Hertz confirmed the existence of an electromagnetic wave by observing the electric current induced by this equipment in a receiving loop (Figure 2.3). Here, there are also two small metallic balls separated by a small gap in the loop. When current is induced in the loop, a spark is generated in the gap.



Figure 2.1 Hertz (left) and Maxwell (right).

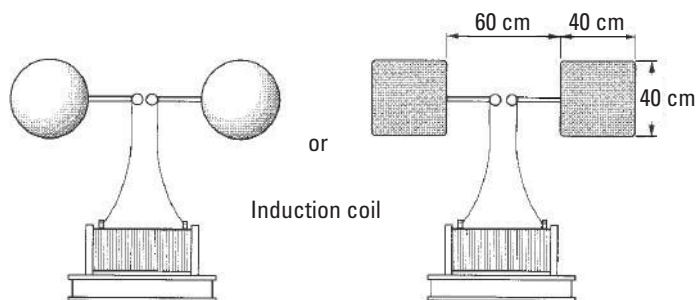


Figure 2.2 Hertz's equipment: an induction coil and two small metallic balls forming a spark gap. The large balls lower the resonant frequency. For an illustrative model, we use square plates (on the right).

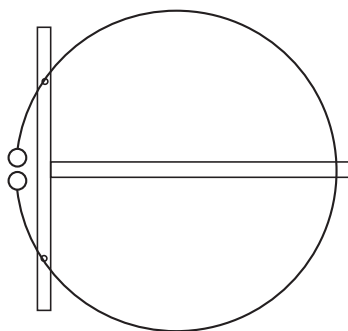


Figure 2.3 Hertz's receiving equipment.

One day, almost by accident, he discovered that he could observe a spark when this device was at some distance from the induction coil shown in Figure 2.2. The loop and spark gap shown in Figure 2.3 corresponds to what is now called a receiver.

2.1.3 Simulation of Hertz's Transmitting Equipment

When we calculate the resonance frequency from the size of the apparatus in Figure 2.2, just like we did for the dipole antenna in Chapter 1, we get $3 \times 10^8 / (1.4 \times 2) = 107$ MHz. Figure 2.4 is a rough model of Hertz's transmitting equipment based on the dipole antenna we created in Chapter 1. This model is composed of two 40-cm square rectangles and two 30×2 -cm thin rectangles. Using this model we can see the nature of Hertz's experiments, which is our present objective. A precise model would require much more complexity and is not considered here.

In the dialog box displayed by selecting Circuit->Units..., set Length Units to cm and Frequency Units to MHz. Next, in the dialog box (Figure 2.5) displayed by selecting Circuit->Box..., set both the x (horizontal, as displayed on your computer screen) and y (vertical, as displayed on your computer screen) Cell Size to 1 cm. Then set the Box Size to 512 by 512 cm as shown. Finally, set Top Metal and Bottom Metal to Free Space, upper right.

Distance from the antenna to the edge of the analysis space is best set to about one wavelength, where the far field is becoming established. However, we set a little less than that in order to stay within the Sonnet Lite memory limit.

The top-to-bottom size of the simulation space is set in the dialog box displayed by selecting Circuit->Dielectric Layers... (Figure 2.6).

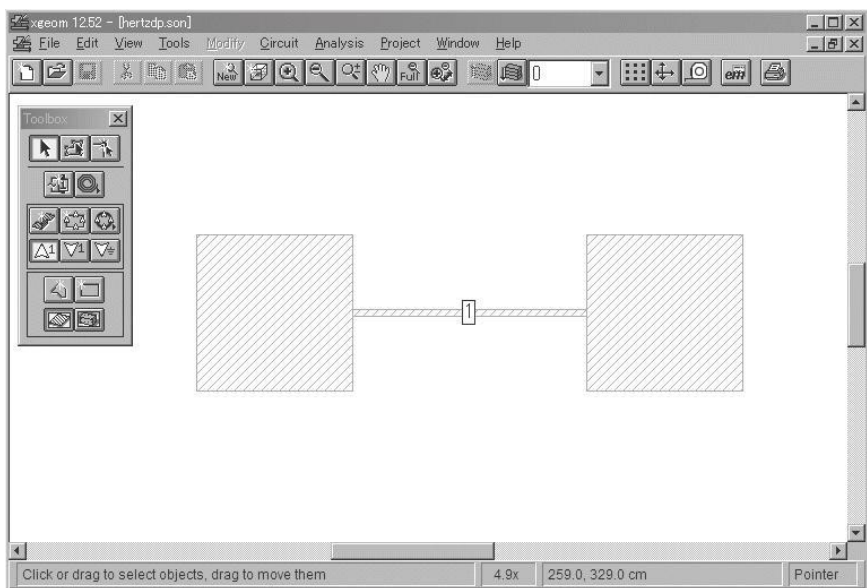


Figure 2.4 Rough model of Hertz's transmitting equipment, similar to the dipole antenna model in Chapter 1.

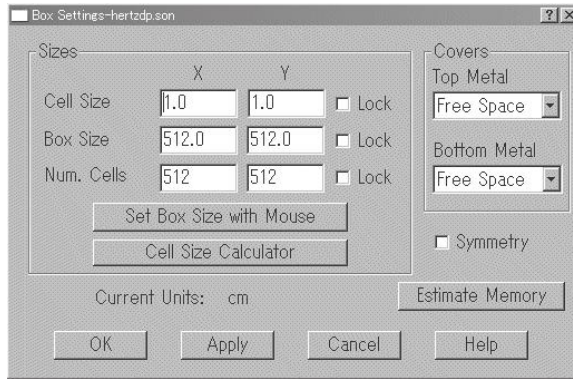


Figure 2.5 Set x (horizontal) and the y (vertical) Cell Size to 1 cm.

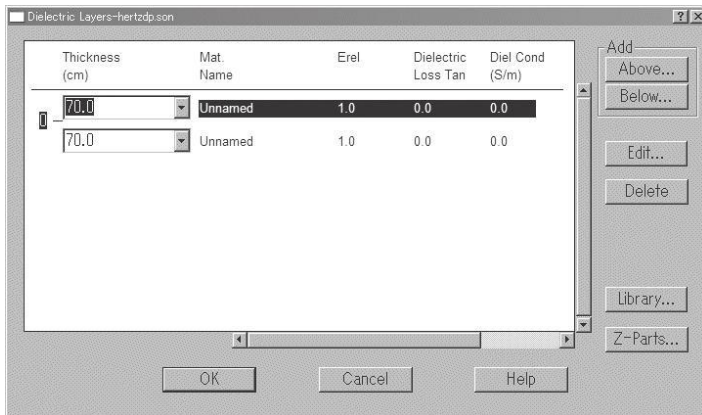


Figure 2.6 The top-to-bottom size of the simulation space is set in this dialog box.

As for the transmitter, we draw rectangles as we did in Chapter 1, and place a port in the center.

Using small metallic balls as a spark gap, Hertz generated high-voltage sparks by interrupting the current flowing through the induction coil. The sparks generate wideband power at high frequency just like the noise from an electric shaver. At that time, there were no transistors, vacuum tubes, oscillators, or amplifiers. Spark gaps were widely used in this manner until the invention of the vacuum tube.

Next, select Analysis->Setup..., set Start to 50 MHz and Stop to 150 MHz. Select Project->Analyze and you will quickly have results.

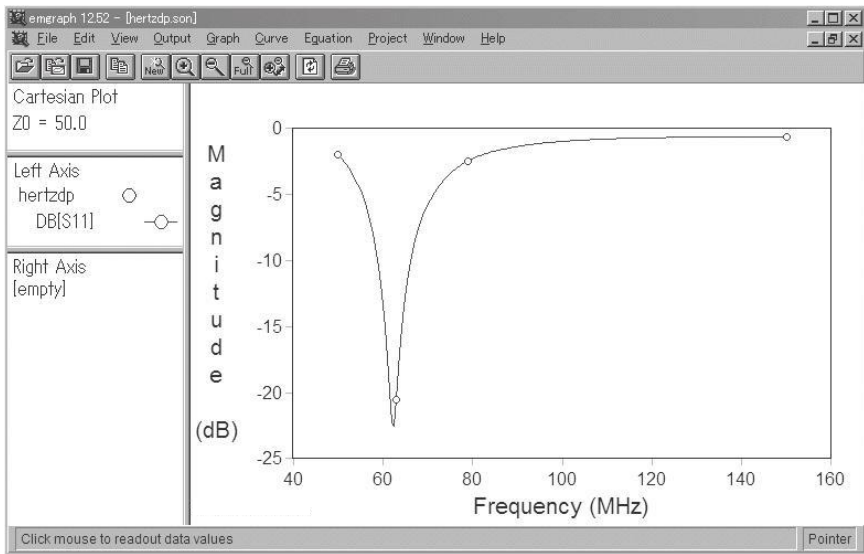


Figure 2.7 Simulation result for S_{11} . The antenna resonates at around 60 MHz, contrary to expectation.

Figure 2.7 is the simulation result for S_{11} , and we see that it appears to resonate at around 60 MHz, which is nearly half the 107 MHz calculated above from the size of the antenna. (A definitive test for resonance requires the input impedance to be purely resistive, as discussed later.) This is due to the large patches on the end of the dipole. This technique of lowering the resonant frequency of a dipole antenna is called capacitive end loading.

Why did Hertz use spheres connected by rods and wires for his antennas? Figure 2.8 shows a Leiden jar used in the 18th century at Leiden University in the Netherlands. Tin foil coats the inside and outside of the glass bottles. A chain drops down from a metallic ball near the top and touches the foil on the inside of each bottle at the bottom, forming one terminal of a capacitor. The foil on the outside of the jar forms the other terminal. Perhaps Hertz's antennas were inspired by this form of the Leiden jar.

Figure 2.9 is the surface charge distribution displayed after selecting Proj->View Charge. Selecting Plot->Response->Current changes the display to charge distribution. We can confirm that a strong current has been stored on the metallic squares at both ends of the antenna at the resonant frequency of 60 MHz.

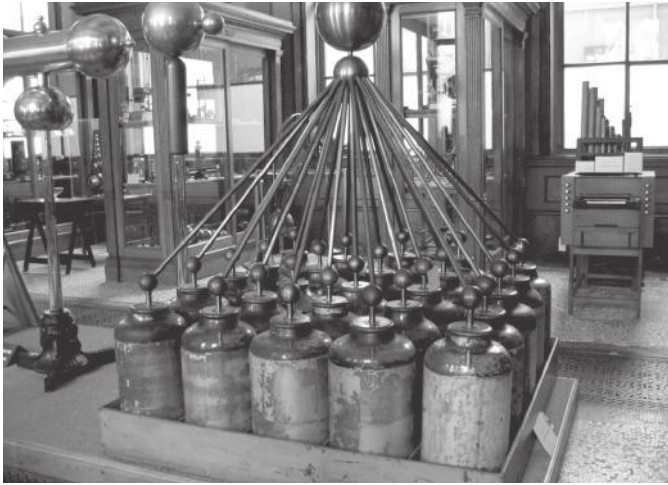


Figure 2.8 The Leiden jar, this one on display at Leiden University, might have been an inspiration for the form of Hertz's antennas.

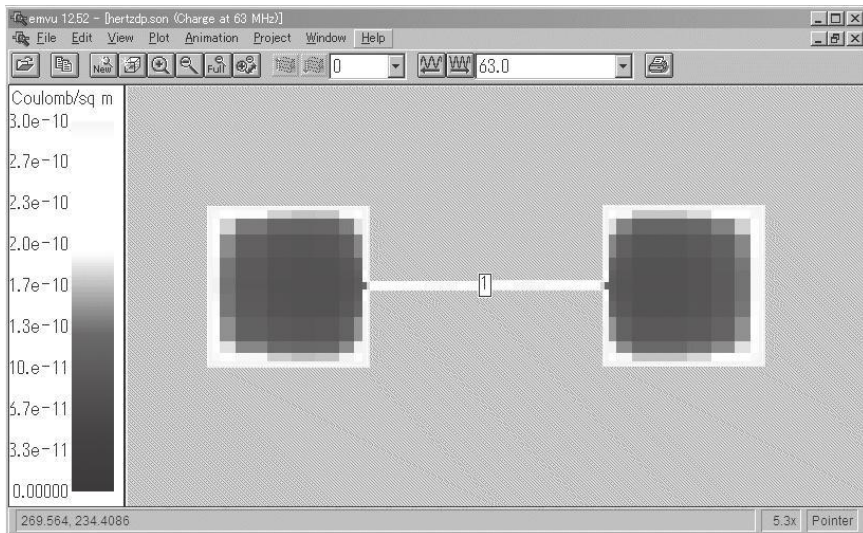


Figure 2.9 Surface charge distribution on the surface of square elements.

2.1.4 Transforming Parallel Plate Capacitors into Antennas

James Clerk Maxwell (1831–1879) (Figure 2.1, right) predicted the existence of electromagnetic waves. We can imagine how Hertz's apparatus generates Max-

well's waves from a thought experiment starting with a parallel plate capacitor. When passing an alternating current through the parallel capacitor of Figure 2.10(a), the voltage across the two plates cause an electric field in the dielectric, as indicated with the dotted lines. There is no radiation. All of the electromagnetic energy stays inside the capacitor.

The capacitor plates become rectangular solids in Figure 2.10(b) with a wider separation. We see that the electric field is distributed over a wider volume. The rectangular solids become spheres in Figure 2.10(c) that are now even farther apart.

In Figure 2.10(a) and (b) where the electric field (electric lines of force) concentrate between the electrodes, there is hardly any radiation. However, we can imagine that when the electric field has extended into space, as seen in Figure 2.8(c), we start to have some radiation, just as with Hertz's transmitter (Figure 2.2). Not shown here is the magnetic field, which also plays an important role, as described below.

2.1.5 Simulation of Hertz's Receiving Equipment

Hertz discovered that the sparks were strongest when the receiving loop was tuned to a specific length (Figure 2.11). Confirming an electromagnetic resonance, with the resonant frequency determined by the electrical dimensions of the receiving loop and the transmitting antenna, was a monumental event. Resonance had only been seen in mechanical systems before.

In a mechanical system, resonance happens when energy oscillates between two different forms. Imagine a pendulum. When the pendulum has swung out as far as it is going to go, it is momentarily stationary. Potential energy is maximum. As the pendulum swings past its central position one-quarter of a cycle later, its velocity, and thus kinetic energy, is maximum. Thus, in reso-

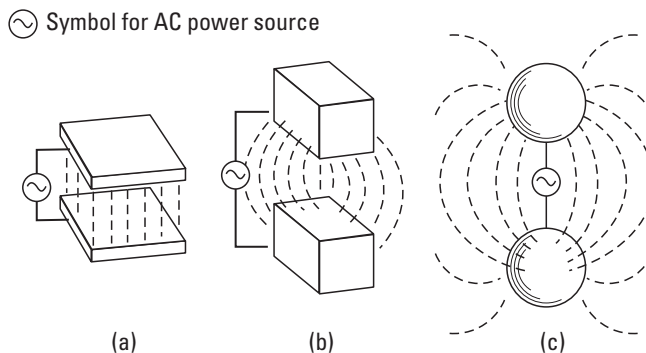


Figure 2.10 (a–c) Electric field (electric lines of force) around the capacitors.

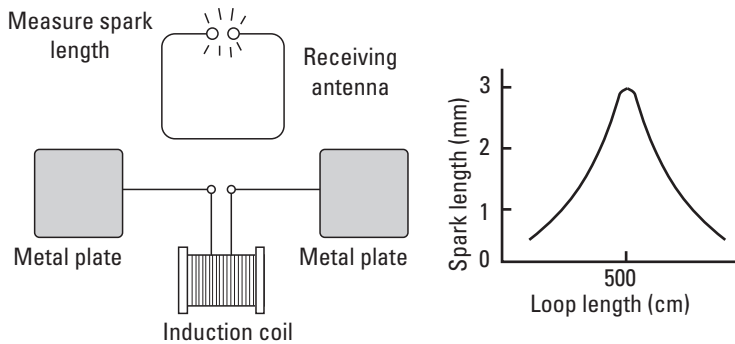


Figure 2.11 Representation of Hertz's 1888 experiment.

nance, energy flows back and forth from one form to another every one-quarter of a cycle.

In an LC (inductor/capacitor) resonant circuit, energy flows back and forth between the capacitor (electric energy) and inductor (magnetic energy). Just like the pendulum, this happens naturally at the resonant frequency.

In Hertz's experimental apparatus (Figure 2.2), the large metallic balls and metallic plates at both sides of the transmitting device are like a large-scale capacitor. The rod that connects them is like an inductor. As for the receiver (Figure 2.3), electric energy (capacitance) is stored between the small balls and magnetic energy is stored around the loop. Electromagnetic resonance happens when energy flows back and forth between electric and magnetic energy. The frequency with which the energy flows back and forth is the resonant frequency.

Figure 2.12 is a simple model of Hertz's receiver in Sonnet Lite. The line width is 2 cm and entire loop length is 500 cm. All other parameters are the same as the transmitter model we analyzed above.

Hertz's receiver also has a gap made with small balls, just like the transmitting equipment. In contrast, the simplified Sonnet model has a port. In Hertz's experiments, an electromagnetic wave generates an output (sparks) in the small receiving loop gap. However, in Sonnet, we must treat the receiving loop as though it were a transmitter. We will input power into the port and, by viewing the S-parameters, see how much is absorbed. Power will be absorbed only when the frequency of the input power is near the resonant frequency.

Figure 2.13 is a plot of S_{11} . We can see that the resonance is near 66 MHz, which is near the 63-MHz resonant frequency of the transmitter. The entire length of the loop of the receiver model is 5m, which corresponds to a frequency of around 60 MHz. So we can see that this loop is about one wavelength long. We see that it looks like there is another resonant frequency on the

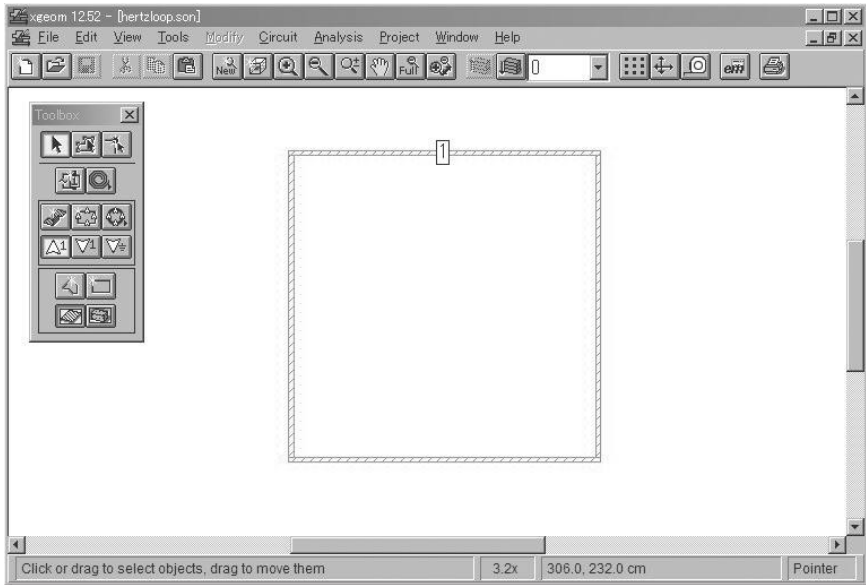


Figure 2.12 Model of Hertz's receiving equipment. Entire loop length is 500 cm.

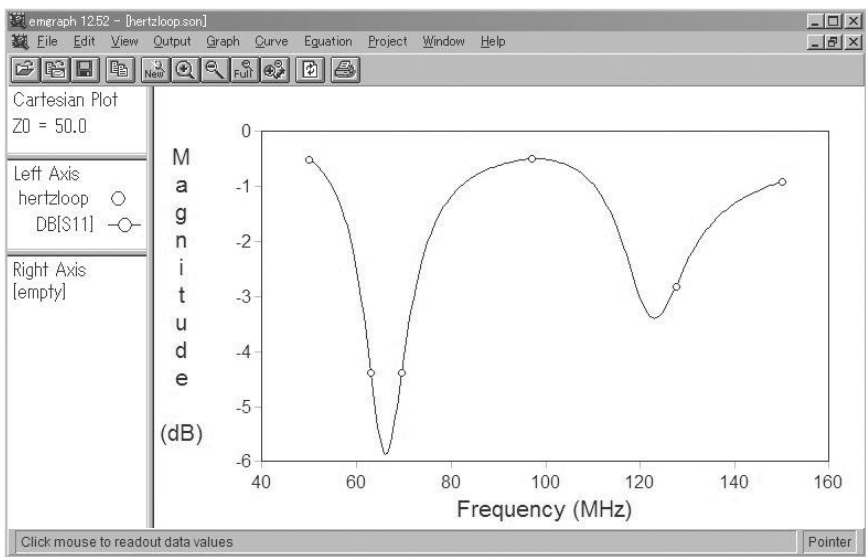


Figure 2.13 S_{11} plot of our model of Hertz's receiving equipment. It resonates at around 66 MHz.

plot at almost double the frequency. The loop is two wavelengths long at the higher resonance.

We see the surface current distribution at 63 MHz in Figure 2.14. One wavelength (two nulls and two maximums) is seen around the entire loop.

With the professional version of Sonnet, we can put both the transmitter and receiver in the same Sonnet analysis. We would leave the transmitter port as port number 1. We would change the receiving loop to port number 2. Then, if we did a plot of S_{21} , we would see how much voltage is present at port 2 given power is being input to port 1.

2.1.6 Experiments by Dr. Nagaoka

In Hertz's time, high-frequency signal generators and frequency counters did not exist. It would have been difficult to correctly anticipate and observe a resonant frequency in the structure in Figure 2.2.

In Japan, the physicist Hantaro Nagaoka did his own experiments just one year after Hertz (Figure 2.15). The results were published in the *Journal of the Science Academy*, No. 7, in 1889.

On page 229 of Nagaoka's work, he wrote, "...when we assume T as the oscillation period, P as the self-inductance, C as the capacitance and V as the ratio of CGS electrostatic units to CGS electromagnetic units, the equation to get T is..."

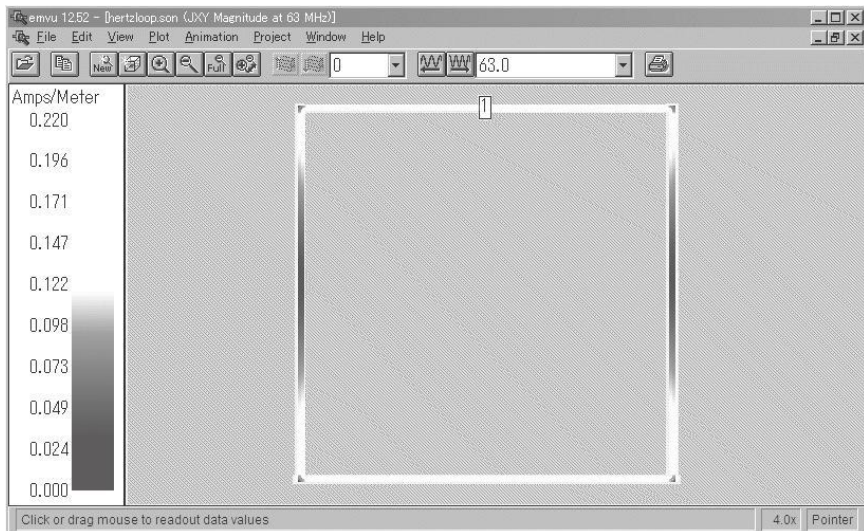


Figure 2.14 Surface current distribution at 63 MHz.

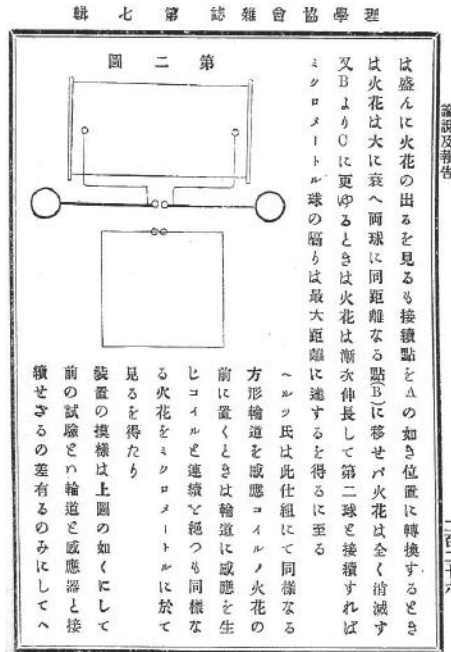


Figure 2.15 “Hertz’s Experiments” by Nagaoka is a vertically written thesis in Japanese.

Dr. Nagaoka’s expression (which was written in an old style of Japanese) corresponds to the resonant frequency f , that is $T = 1/f = 2\pi\sqrt{LC}$ (using modern notation). Subsequently, he obtained the value of P (L in the equation above) from the length and radius of the transmitting antenna and the value of C from the radius of the metallic ball, to calculate the resonant frequency.

This thesis was based directly on Hertz’s publications, and we can read his description “...Hertz also explored how he could get the longest spark discharge with the help of resonance; he finally discovered a convenient way to study electromagnetic waves.”

According to this description, Hertz determined the resonant frequency of his transmitter by calculation. Hertz also checked several lengths of receiving loop to verify a resonance condition.

2.2 The Development of Various Antennas

Hertz was convinced that his discoveries were of absolutely no practical use. But the foundation that he laid was now in place for others to use.

2.2.1 Appearance of Marconi

In the history of cable communication, Samuel Morse, who is famous for the Morse code, laid the very first submarine cable in 1842. During his experiments, the anchor of a ship happened to cut the wire right in the middle of his first public demonstration. Because of this embarrassing episode, he started experimenting with a “wireless” telegraph, using earth and bodies of water as electrical conductors. His experiments were successful in communicating over a mile or so across several rivers.

Guglielmo Marconi (1874–1937) in Italy had learned about Hertz’s experiments. He reproduced some of Hertz’s experiments and then he started using the transmitters and receivers for communication. He established Marconi Corporation to commercialize his work.

Figure 2.16 is a chronological table that shows the evolution of grounded antennas. Marconi’s antennas and communication equipment are listed first. He successfully carried out communication over a distance of 2,400m with an antenna 8m high. Its main feature was use of a wire to earth to ground the globe. The end of the antenna was grounded to the earth and the transmitter was connected between ground and the antenna.

The transmitter (Figure 2.17, left, on the desk) is composed of four globes and Hertz’s inductive coil (right).

A copper plate is suspended near the top, which corresponds to the metallic ball on one side of Hertz’s antenna. Under the desk, we can see a horizontally oriented copper plate of the same size.

This metallic plate is much larger than Hertz’s spheres. Marconi was intensely focused on extending the communication distance and had experimentally determined that he could transmit farther with a larger metallic plate.

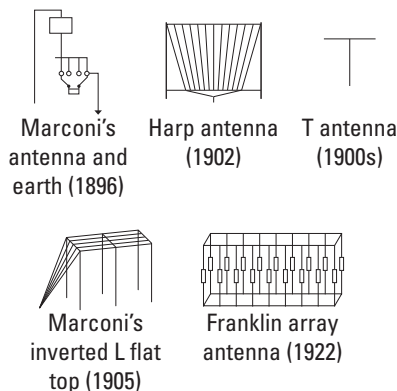


Figure 2.16 Evolution of grounded antennas.

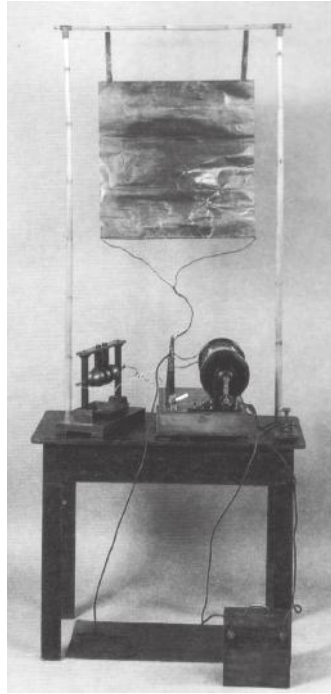


Figure 2.17 Marconi's transmitter.

According to Dr. Nagaoka, the metallic plate is capacitive. So if the plate is larger, the resonant frequency is lowered. Marconi would have noticed that a longer wavelength would transmit further. As he made his antenna capacitance larger and larger, he hit on the idea to use a globe as the opposite capacitance.

The harp antenna in Figure 2.16 is also one of the antennas that Marconi built. When his 60m tall antenna was blown down in a gale, he rebuilt it in the shape of a harp with the height of 45m to carry out the experiments, successfully communicating over a of 360-km path.

2.2.2 History of Ungrounded Antennas

Figure 2.18 shows the evolution of ungrounded antennas. The ancestor of these antennas is Hertz's oscillator. Next came the resonant antenna that physicist Oliver Lodge in Britain (1851–1940) invented. He changed the metallic plate (metallic ball) of Hertz's oscillator to a cone shape, which is connected to the coil at the center. As Hertz had determined, the size of the capacitor and the number of turns of the coil can be changed, and resonant frequency adjusted to desired values.

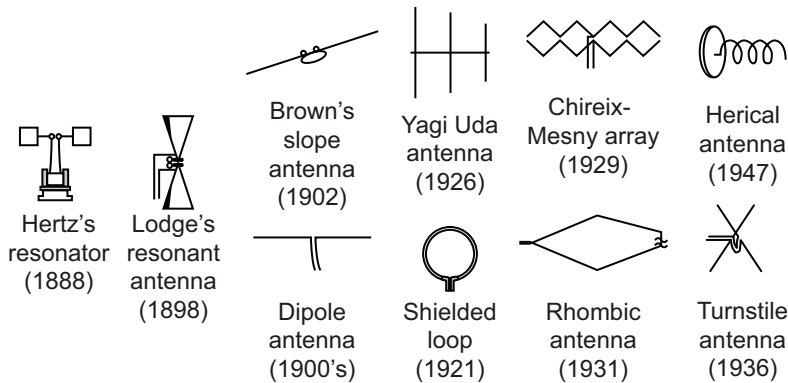


Figure 2.18 Evolution of ungrounded antennas.

Both grounded and ungrounded antennas were complex structures in the early days. However, the dipole antenna was simplified to a metallic rod in the 1900s. This simple structure has the capacitive body at both ends of Hertz's dipole removed because it was discovered that electromagnetic waves could be transmitted and received just as well. Today, electronic systems are becoming more and more complicated. However, for antennas, at least in the past, the reverse has sometimes happened.

The Yagi-Uda antenna in the middle of Figure 2.18 is the world-famous Yagi antenna invented in Japan. In spite of its simple structure with dipole antennas of slightly different length mounted on a central boom, it can transmit and receive electromagnetic waves very well in a desired direction, while also attenuating waves from other directions. As such, it can be used as a narrow band (i.e., highly optimized, single channel) receiving antenna for television. It is typically not used as a broadcast TV or FM antenna, as broadcasters like to transmit their signals in all directions.

2.2.3 Aperture Antennas

Figure 2.19 shows the evolution of aperture antennas. They have an area, or aperture, opening towards the desired transmit/receive direction. For example, the horn antenna in the middle looks exactly like a horn. It functions very much like a megaphone, focusing the transmitted signal in the desired direction. The horn is made of metal.

The ancestor of this antenna type is also due to Hertz. To confirm that the electromagnetic waves he had generated have the behavior of light, he placed a dipole antenna at the focus of a parabolic reflector formed from wires (Figure 2.20). This is analogous to the parabolic mirror in a reflecting telescope

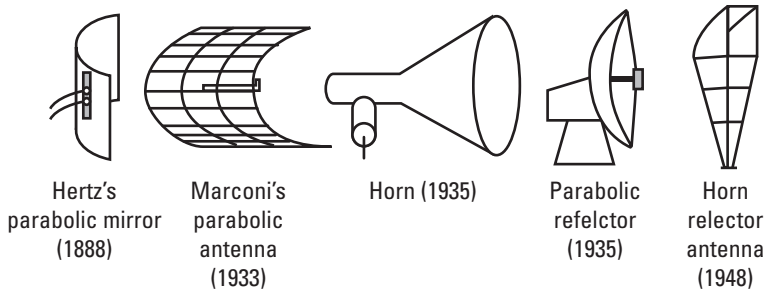


Figure 2.19 Evolution of aperture antennas.

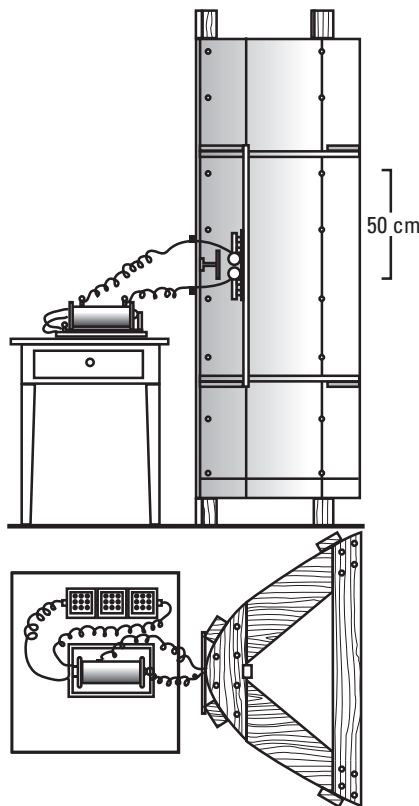


Figure 2.20 Hertz's transmitting equipment with a parabolic reflector.

focusing and concentrating light. After all, it was thought, light must also be an electromagnetic wave.

Hertz confirmed that he could receive waves even if he put the receiver 20m away, at a wavelength of 66 cm. He also discovered that no sparks were observed when the transmitter antenna and receiver antenna were at right angles (say, one horizontal and the other vertical). Electromagnetic waves were polarized, exactly as predicted by Clerk Maxwell.

Once it had been discovered that light is polarized, they knew that light could not be a longitudinal wave, like sound. Sound is a pressure wave. It has only one direction of vibration: along the direction it is traveling. If light is polarized, then it must be vibrating side-to-side, perpendicular to the direction it is traveling. This is called a transverse wave.

The direction of polarization for an electromagnetic wave refers to the direction of vibration of the electric field. When placing a dipole antenna vertically above ground, the electric field vector is also vertical, and we call it a vertical polarization. Marconi's antennas were vertically polarized.

When we place an antenna horizontally above ground, it generates horizontally polarized waves. Hertz's transmitter shown in Figure 2.2 is horizontally polarized.

The horn antenna in Figure 2.19 includes a small antenna inside and at the left end of the horn, which is fed by coax. The small antenna launches an electromagnetic wave that is guided down the length of the horn. As the horn gradually widens, it gradually transitions the wave to free space. The longer and more gradual the widening of the horn, the more directive the antenna, just like a megaphone.

As we can see in the evolution charts (Figures 2.16, 2.18, and 2.19), there are many types of antennas. Some of these, including built-in cell phone antennas, are becoming more widespread and often need to be miniaturized as much as possible.

2.2.4 The Role of Ground

The pair of antennas for WiMAX described in Chapter 1 has a ground conductor located between them that is an integral part of the antenna. The ground that we learn about in textbooks is an ideal conductor, and is also called an electric wall in electromagnetic field theory. When we have an ideal conductor extending over a large area, it acts just like a mirror, so such an antenna can be viewed as a ground-image antenna (Figure 2.21). One half of a dipole mounted above a large ground acts almost exactly like a complete dipole antenna that has no ground at all.

The current distribution that is drawn for the half-dipole that is above the ground is the current that actually exists on the half-dipole. Notice that current is also drawn below ground. This current does not really exist. However, electri-

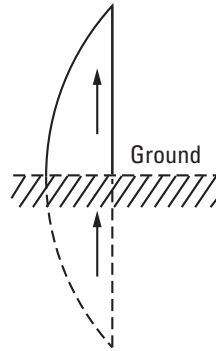


Figure 2.21 An image antenna.

cally, the half-dipole plus ground plane acts almost exactly like a full dipole with no ground plane.

To illustrate this, we now modify Hertz's dipole and run a simulation for Marconi's antenna. First, load the previous dipole model and delete the entire right side. To do this, click your mouse on the desired element and press the Delete key. If Port 1 disappears, just add it back in.

Next, left-click and drag a rectangle around the entire remaining antenna, then select Modify->Rotate... and rotate it clockwise by 90 degrees (Figure 2.22). Instead of left-clicking and dragging to select the entire circuit, you can also type CTRL-A.

As for the Sonnet simulation space, the four substrate edges each mark one of four perfectly conducting side walls. To create a ground-image antenna, select and drag the entire antenna and place it so that Port 1 just touches the

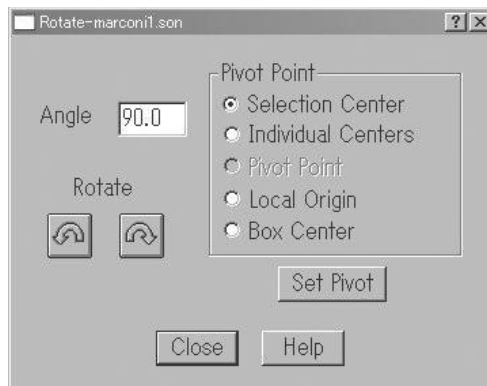


Figure 2.22 Dialog box to rotate the antenna.

lower edge of the substrate (Figure 2.23). The edge of the substrate marks a perfectly conducting sidewall, making this a ground-image antenna.

Previously, port 1 excited a voltage between the two halves of the dipole. Now, that same port is exciting a voltage between the half dipole and the ground plane. An antenna like this is called a *monopole*.

Figure 2.24 is a plot of S_{11} showing results of both the monopole and the dipole. Even though the antennas are different, the resonant frequencies are almost the same. However, the plots are different. The monopole has higher reflection, more power is reflected and less is radiated. Perhaps the input impedance has changed?

Figure 2.25 shows the resistance, R , of the input impedance. You can see the value of the monopole antenna is approximately half of that of the dipole.

The textbook input impedance of a dipole antenna is about 73Ω pure real (no reactance). In contrast a monopole is 36Ω , which is just half. Why is this? Figure 2.26 shows a monopole antenna and the equivalent dipole antenna.

Both antennas have the same current. However, the voltage at the feed point is V for monopole antenna and $2V$ for dipole antenna. Impedance is the ratio of voltage to current, so the input impedance of a monopole antenna is half that of a dipole antenna.

Keep in mind that we are assuming the ground is an ideal conductor and it is infinite. As for a finite ground (remember the WiMAX antenna of Figure

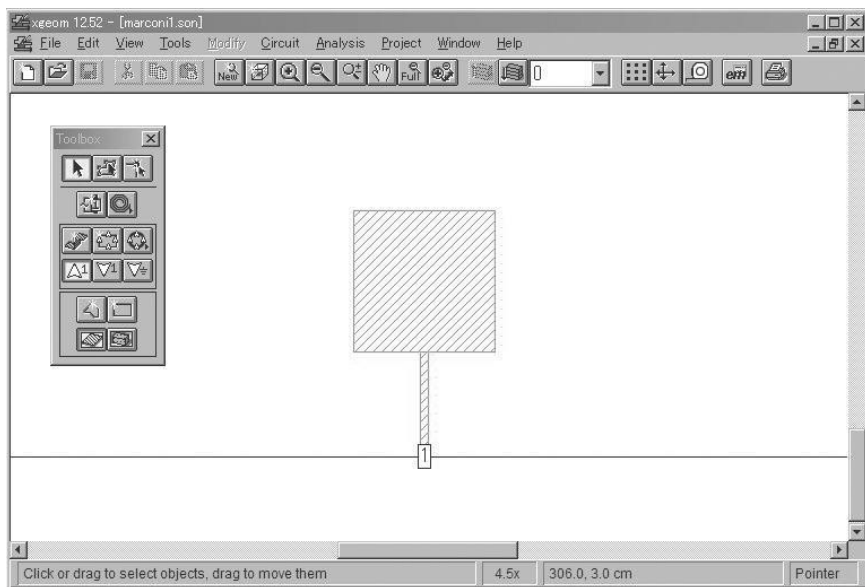


Figure 2.23 Move the entire antenna so that port 1 just touches the lower substrate edge (side wall).

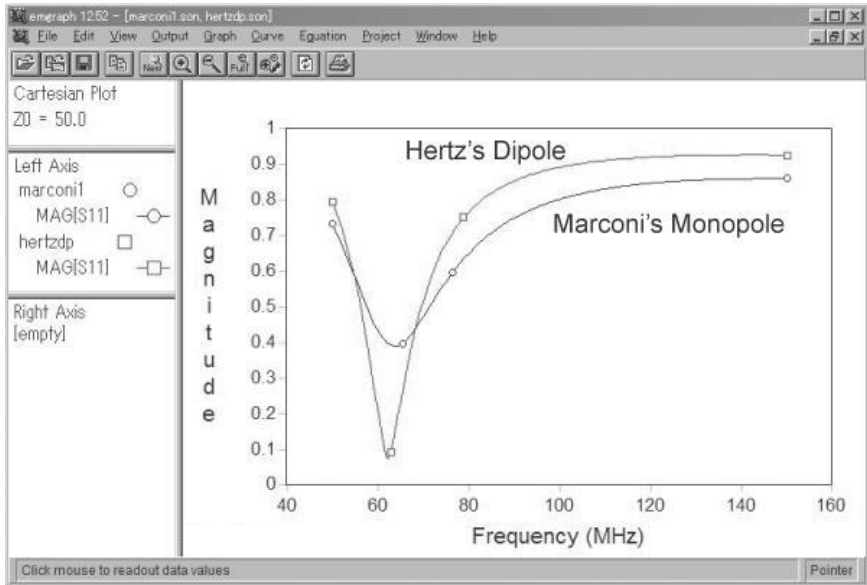


Figure 2.24 S_{11} of Marconi's monopole and Hertz's dipole.

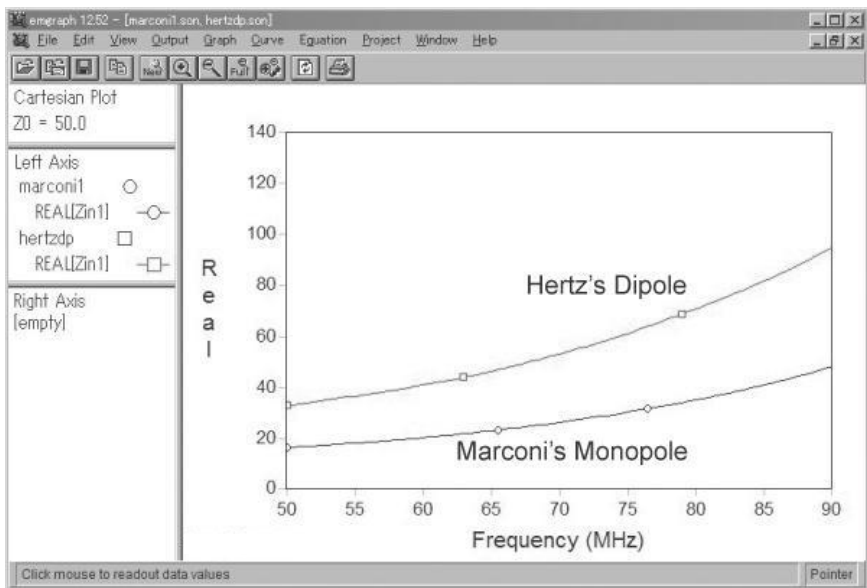


Figure 2.25 Input resistance of Marconi's monopole and Hertz's dipole.

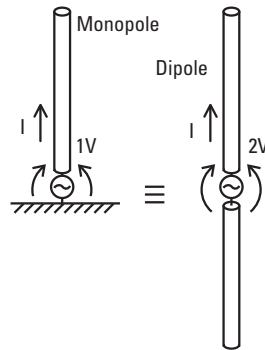


Figure 2.26 A monopole antenna is the equivalent of a dipole antenna.

1.3), the input impedance of an antenna depends on the size of the antenna and on the size of the ground.

2.2.5 Current on an Artificial Ground

Figure 2.27 shows the result when feeding Marconi's antenna with an artificial ground of large metallic plate (Sonnet Lite cannot run this example because of memory limits).

2.3 Electric Field, Magnetic Field, Electromagnetic Field, and the Electromagnetic Wave

Fields and waves; let's make sure we know how they are all related.

2.3.1 Electric Field Near Hertz's Dipole

Figure 2.28 (left) shows the charge distribution of Hertz's dipole at a fixed point in time. It called a dipole because electric field lines go between two (di) poles, one plus pole and the other minus. In later years, it was discovered that it still works as an antenna even if Hertz's metallic balls are taken away and ordinary wire is used (Figure 2.25, right).

2.3.2 Radiation from a Dipole Antenna

Figure 2.28 shows the near field around Hertz's dipole. Now let's think about the electric field in the far field extending beyond it. When applying a high-frequency voltage to Hertz's dipole, the electric lines of force vary as shown in Figure 2.29. Here, the fields of one cycle are drawn, and we can see how the

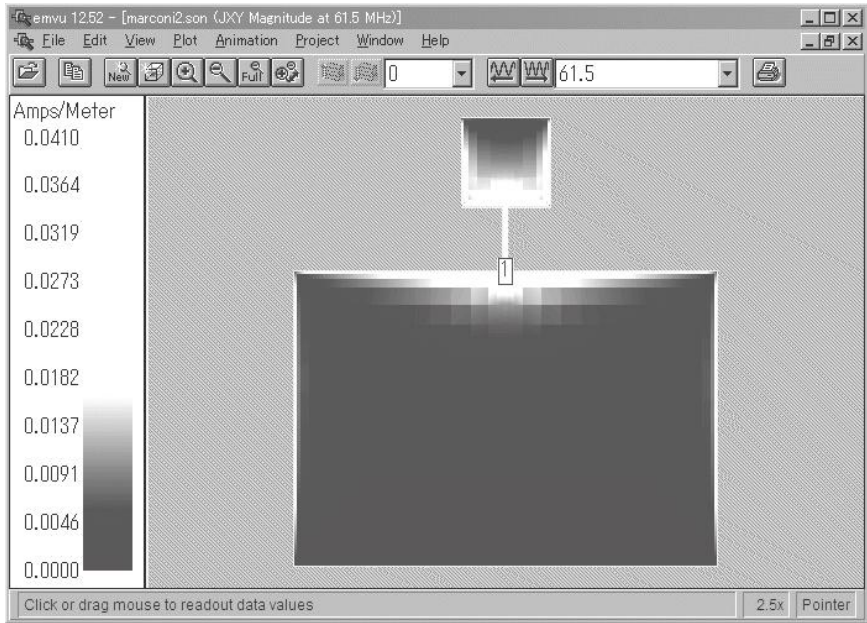


Figure 2.27 Simulated current on Marconi's monopole and an artificial ground.

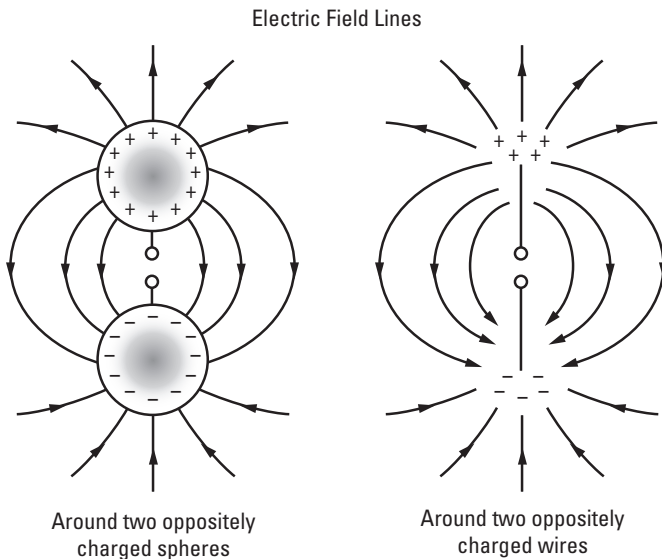


Figure 2.28 Near field around Hertz's dipole and a dipole made of metallic wire. All quantities are varying sinusoidally with time; they are illustrated for a single instant.

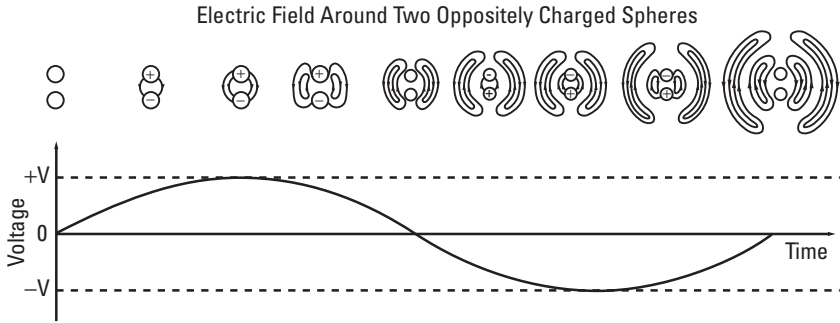


Figure 2.29 Electric lines of force around Hertz's dipole propagate into the distance.

electric lines of force between the positive and negative charges are radiated into space like puffs of smoke. The loops of these electric lines of force stretch out over larger and larger distances as time goes on and expand into space. We can imagine that the electric energy is propagating into the distance.

2.3.3 Magnetic Field Near the Dipole Antenna

Figure 2.30 shows the magnetic field generated around Hertz's dipole. In the near field, we can observe magnetic lines of force according to Ampere's law (one of Maxwell's equations), which propagate into the far field.

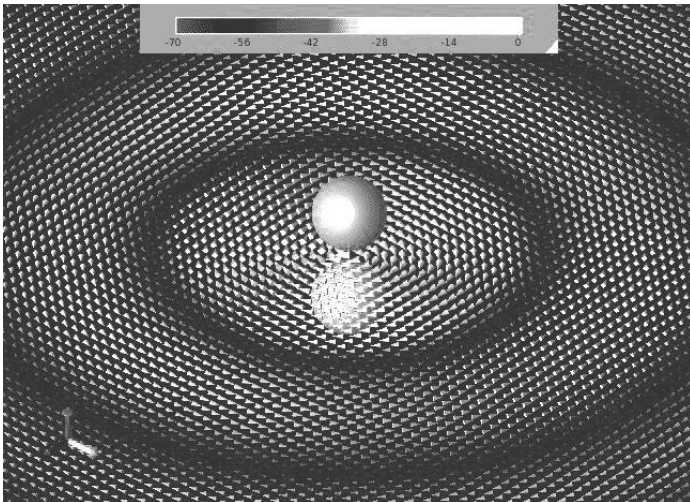


Figure 2.30 Magnetic field generated around Hertz's dipole.

Magnetic lines of force are directed alternately clockwise and counter-clockwise. In the transition region, where the direction changes, the magnetic fields are zero (black in Figure 2.30). Realizing that we are dealing with an electromagnetic wave, we see that the point where magnetic field is zero appears every half-wavelength (XFDTD7 was used for these results).

2.3.4 Electromagnetic Field and the Electromagnetic Wave

Faraday's law of induction, described in Chapter 1, states that "a time-varying magnetic field generates an electric field." Combining this rule with the rule that "the time-varying electric field generates a magnetic field" that Maxwell discovered, the time-varying electric field generated by Hertz's dipole makes a time-varying magnetic field, which in turn also makes a time-varying electric field. So the electric field and the magnetic field travel in space as a wave, each regenerating the other as they both vary with time. The mutually time-varying electric and magnetic fields propagate with the speed of light, and we have an electromagnetic wave.

In Japan, an electromagnetic wave is sometimes simply called an electric wave. In Japanese radio law, a radio wave is an electromagnetic wave below 3 THz.

Clerk Maxwell predicted the existence of an electromagnetic wave and he then calculated the velocity of an electromagnetic wave traveling in the space. Its value is approximately 3×10^8 m/s, which he knew seemed to be the same as that of light. He then proposed in his electromagnetic theory of light (1864) that in fact light is an electromagnetic wave. He did not consider the possibility of lower frequency electromagnetic waves.

There are two kinds of electric charge, plus and minus. When we have a charge of any sign, the direction of force on another nearby positive charge is defined as the direction of electric field. The intensity of the electric field is stronger where the force is stronger. Thus, the electric field is a vector, which has an intensity (i.e., magnitude) and direction.

The electric and magnetic fields vary with time. For example, when no one is talking, the electric and magnetic fields close to an AM or FM radio station antenna are a perfect sine wave at a given frequency. However, the sine wave for the electric field is slightly different from the sine wave for the magnetic field. Specifically, the phase between the electric field and magnetic fields differ by 90 degrees. If the electric field is a sine wave, this means that the magnetic field is actually a cosine wave, a sine wave delayed by 90 degrees. This is the nature of electromagnetic resonance, it can be imagined that in the near field, the electric energy and the magnetic energy are sloshing back and forth. First all the energy is in one, then, one-quarter of a cycle later, all the energy has

surged into the other. So if all the electromagnetic energy is just sloshing back and forth, how does it ever travel to the distant receiver?

Maxwell's equations also give us a traveling wave, shown in Figure 2.31. For this figure, we are showing the electric and magnetic field only along a single line along the direction of travel. The complete electric and magnetic fields are everywhere the same on a transverse (i.e., perpendicular to the direction of travel) plane. Note that the electric field E_x and magnetic field H_y are perpendicular to each other and are in phase. We call this a plane wave.

If you are familiar with the vector cross product, $\mathbf{E} \times \mathbf{H}$ is interesting. It has units of W/m^2 because it is the product of the electric field (V/m) and magnetic field (A/m). We call it the Poynting vector. It is named after the professor who first discovered it. We can think of it as the direction and amount of power flowing through a unit area. It is just an accident of history that the name of this vector makes it easy to remember that it “points” to where the power is flowing.

In the near field of an antenna, the electric and magnetic energy surge back and forth, going nowhere. In the traveling wave, the electric and magnetic fields are in phase, traveling at the speed of light. Radiation happens as the electric and magnetic fields gradually start to oscillate in phase in the transition between the near and far fields. All of this is found, mathematically, in Maxwell's equations. Perhaps you can start to see why engineers who have worked with Maxwell's equations all their lives can consider them to be the most beautiful poetry ever written.

Electric field (electric lines of force) and magnetic field (magnetic lines of force) cannot be seen or experienced directly, but we can see how Maxwell, Hertz, and Marconi might have imagined electromagnetic waves back in those days.

2.3.5 Difficulty of Near-Field Problems

Figure 2.32 shows an antenna field distribution (from XFDTD7). At resonance, electric energy and magnetic energy alternate every one-quarter cycle, going

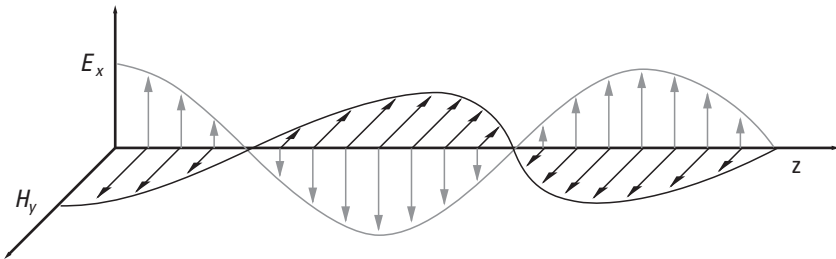


Figure 2.31 Plane wave traveling in space. The electric field vector and magnetic field vector are orthogonal (i.e., perpendicular).

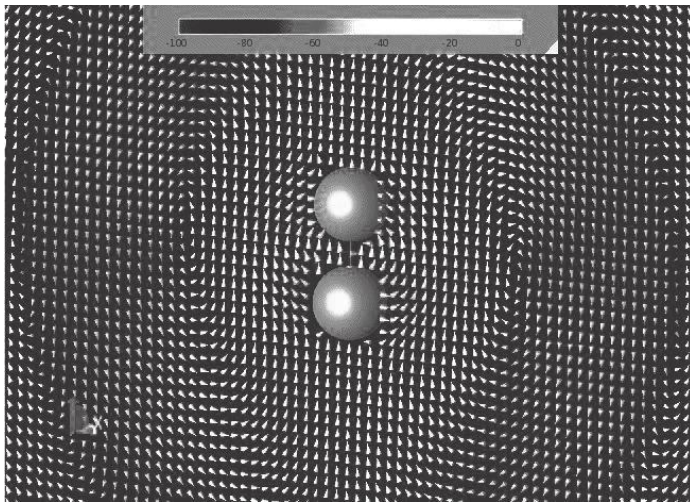


Figure 2.32 Electric field near Hertz's dipole. A loop of the E field can be seen at the border between near field and far field.

nowhere. In the far field, the electric and magnetic energy are in phase, and are mathematically simple. The electromagnetic field distribution in the near field is complicated. For that reason, in the past, engineers sometimes try to avoid working with the near field.

2.4 Antenna Design by Using EM Simulators

EM simulators now allow engineers to deal with the near-field complexity in detail. In this section, we demonstrate how that is done.

2.4.1 Antennas on PCB

It has been said that the antenna is the ultimate electromagnetic field problem. We need to solve Maxwell's equations (Figure 2.33) precisely. Circuit designs using Ohm's law can be calculated with a handheld calculator. In contrast, the vector calculus operations for antenna designs are difficult to do by hand, and then, only for a few simple antennas. So a computer is usually needed. Fortunately, we can use a free electromagnetic simulator to master antenna design techniques as we travel the path to success.

Once you have learned the techniques, there are a wide variety of EM computer tools you can use to design antennas. However, we use Sonnet Lite because it is both easy to use and free. Sonnet Lite does EM analysis of antennas on a planar substrates. This software is a free version of the full professional

$\text{rot} \mathbf{E} = - \frac{\partial \mathbf{B}}{\partial t}$ $\text{rot} \mathbf{H} = \mathbf{J} + \frac{\partial \mathbf{D}}{\partial t}$ $\text{div} \mathbf{B} = 0$ $\text{div} \mathbf{D} = \rho$	or	$\nabla \times \mathbf{E} = - \frac{\partial \mathbf{B}}{\partial t}$ $\nabla \times \mathbf{H} = \mathbf{J} + \frac{\partial \mathbf{D}}{\partial t}$ $\nabla \cdot \mathbf{B} = 0$ $\nabla \cdot \mathbf{D} = \rho$	Faraday's law Ampere's law and Maxwell's displacement current No magnetic charge Coulomb's law
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Maxwell's equations

Figure 2.33 Maxwell's equations.

version of Sonnet Suites and is based on the method of moments, originally written by one of our authors, Dr. James C. Rautio. Thanks to amateur radio, the three authors of this book have been friends for a very long time. Installation instructions for Sonnet Lite are in the appendix. The software is in the attached DVD and can also be downloaded from www.sonnetsoftware.com.

Sonnet is an electromagnetic simulator of planar 3-D structures. Basically, we draw the antenna in Sonnet, specify one or more layered substrates, and then we analyze.

2.4.2 Antennas Created by Using EM Simulators

Modern antennas differ vastly from classic types like the dipole and Yagi antennas. Today, they are very compact and often embedded in the product. Older cell phones had an external wirelike antenna. Now it is standard practice to embed the antenna inside the housing.

Here, we learn new design procedures that are necessary to develop compact antennas on a substrate. First, we summarize the antenna types that we have simulated up to now.

1. Linear dipole antennas (Chapter 1, Section 1.3);
2. Yagi antenna with two elements (Chapter 1, Section 1.3);
3. Hertz's transmitter (dipole antenna) (Chapter 2, Section 2.1);
4. Hertz's receiver (loop antenna) (Chapter 2, Section 2.1);
5. Marconi's transmitter (monopole antenna) (Chapter 2, Section 2.2).

These fundamental types are all electric field detection antennas and 1 through 4 are antennas for which the ground is not an integral part; antenna 5 is a ground-image antenna. You might think 4 is the magnetic field detection type because it is a loop antenna. However the entire length is one wavelength and it is just like two half-wavelength dipoles connected end-to-end, so we describe it as an electric field detection type.

As described in Chapter 1, the magnetic field detection antenna is small, one-tenth of a wavelength or less, and its design procedure is different from the electric field detection antenna. We simulate a small loop antenna at the end of this chapter to demonstrate.

2.4.3 Design of Electric Field Detection Type Antennas

The design procedure of a dipole antenna is:

1. Determine the length of one-half wavelength at the operating frequency. For a straight dipole in air, assign an initial length of about 97% of one-half wavelength. End capacitance requires that the physical length of a dipole be slightly shorter than a one-half wavelength. Any substrate material requires the dipole to be even shorter.
2. In Sonnet, draw the dipole in two equal pieces. Place a port in the center where the two pieces meet.
3. Set the simulation volume so that the antenna is one wavelength (or more) away from all sidewalls (substrate edges).
4. Put the antenna about one-half a wavelength away from the Top and Bottom covers of the Sonnet simulation space by setting the appropriate layer thicknesses.
5. Set the Top and Bottom covers to Free Space.
6. Set the desired frequency range in an ABS sweep. Alternatively, you can also set Linear Frequency Sweep if you do not want interpolated results (this will take longer to analyze).
7. Analyze and confirm the resonant frequency by viewing S_{11} .
8. Confirm the input impedance. For a straight dipole in air, it must be close to 73Ω pure real.

Detailed descriptions of how to do the above steps are illustrated in the examples of Chapter 1. Meander antenna elements like a WiMAX antenna (Figure 1.3) do not have the same length as straight dipole antenna, so it is necessary to repeat the above procedure to find the suitable size and shape. For

example, if the apparent analyzed resonant frequency is too high, you must make the antenna longer.

Refer to Chapter 8 for more simulation examples that you can try for yourself.

2.4.4 Simulation of a Small Loop for an Integrated Circuit Tag Antenna

Figure 2.34 shows a one-turn coil that is about the size of a credit card. This antenna (or, more accurately, coil or inductor, because it is a tiny fraction of a wavelength in size) is different from most of the antennas we have worked with so far. There is no coax or feed line. Instead, it is connected to an integrated circuit (IC) that feeds the antenna directly. Coils of a few turns are used for the cards like this, for example, FeliCa™, developed by SONY.

Here we simulate a one-turn coil using Sonnet Lite. To stay within Sonnet Lite's memory constraint of 16 Mb, set the cell size in both the x and y directions to 2 mm. The width of the metal line forming the coil is also 2 mm, so the coil line will be meshed one cell across. With the full Sonnet professional version, you can choose a smaller cell width and get more accurate results. This would be important, for example, if you need high accuracy for calculation of loss.

We assume the coil is made of aluminum. After selecting Circuit->Metal Types... in the displayed dialog box, click the Add... button and then click

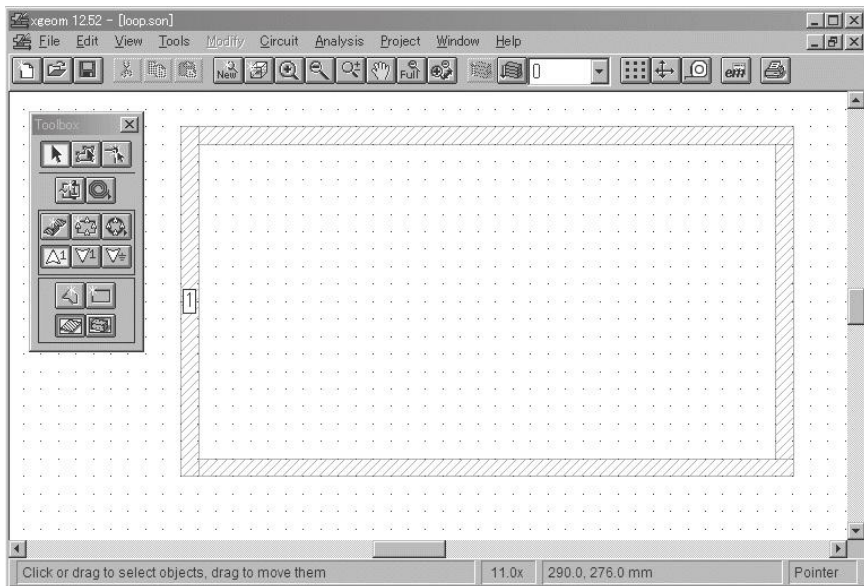


Figure 2.34 A one-turn coil of credit card size. Model file: loop.son.

Select metal from library..., and then, in the Global Library, select Aluminum (Figure 2.35).

The value of Conductivity is entered automatically. Assign 0.01 mm (10 μm) to Thickness. Next, set the Current Ratio to 1, which means we are assuming equal current on the top and bottom surfaces of the metal foil.

As for Circuit->Box, set both the x and y dimensions of the substrate surface to 512 mm. Then, in Circuit->Dielectric Layers, set the top-to-bottom distance to 70 mm (two layers of air, each 35 mm thick). This tag's operating frequency is 13.56 MHz, so the wavelength is about 22m. However, this antenna uses only the magnetic field generated near the coil, so we simply set the Sonnet simulation space large enough so it does not interfere with the coil's magnetic field.

Drawing the coil is similar to drawing Hertz's receiving loop antenna, so we set Port 1 as we did with Hertz's antenna. This IC card has aluminum foil on the surface of polyethylene terephthalate (PET, a kind of polyester) film, so add another dielectric layer. After selecting Circuit->Dielectric Layers..., Add Layer, set the Thickness to 0.05 mm (50 micron), relative permittivity to 2.2, and Dielectric Loss Tan to 0.001 (Figure 2.36).

After selecting Analysis->Setup..., select Frequency Sweep Combinations. Then select Single Frequency from the drop-down menu and click the Add... button to input 13.56 MHz. Next, select Adaptive Sweep (ABS) from the drop-down menu and then click the Add... button, input 13 MHz and 14 MHz. Select Project->Analyze to run the simulation.

After displaying S_{11} , select Equation->Add Equation Curve. Select Inductance1(nH) and click the OK button. Sonnet now displays a plot of inductance. According to this plot, L is 167.2 nH (nanohenry). Resonant fre-

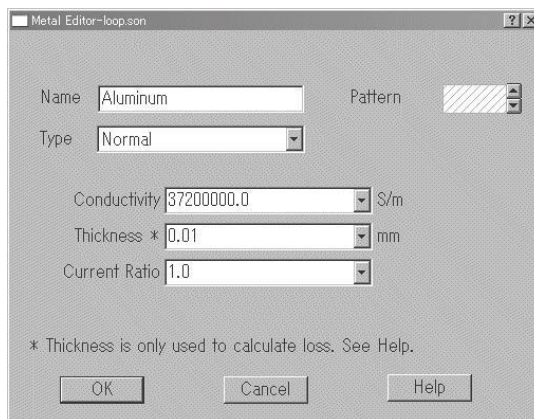


Figure 2.35 Making aluminum available for selection.

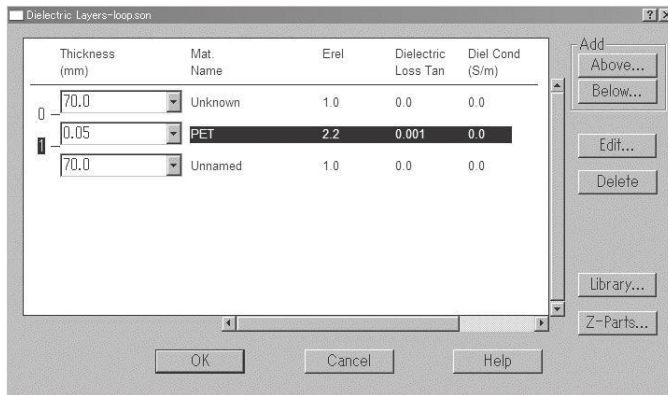


Figure 2.36 Add another dielectric layer to model the PET film upon which the coil will be placed.

quency is $f = 1/2\pi\sqrt{LC}$. A little algebra gives us the C to resonate at 13.56 MHz: 823.9 pF.

To confirm that that value of capacitance creates the desired resonance, we connect a capacitor across the coil to make an LC resonant circuit. To do this, we create a Sonnet netlist. Select File->New Netlist in the main Sonnet window and the window shown in Figure 2.37 appears. In our initial display of this window, it is set for wiring up a 2-port. We want to analyze a 1-port. So double click on that one line in the net list and change # of ports from 2 to 1 and click OK.

Next, selecting Tools->Add Project Element..., and assign the file name loop.son. This is the project we just finished simulating. Now select Tools->Add Modeled Element..., and add the capacitor. Select Analysis->Setup... and assign the same frequency sweep as in the previous project. Then select Project->Analyze.

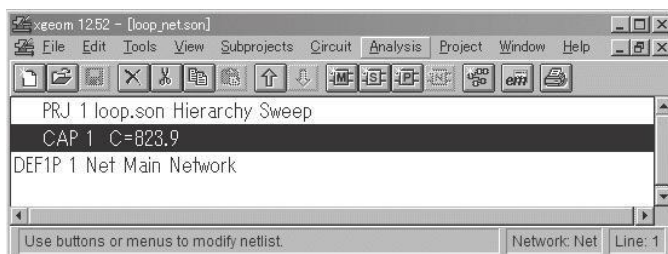


Figure 2.37 Final window for the Sonnet Netlist Project.

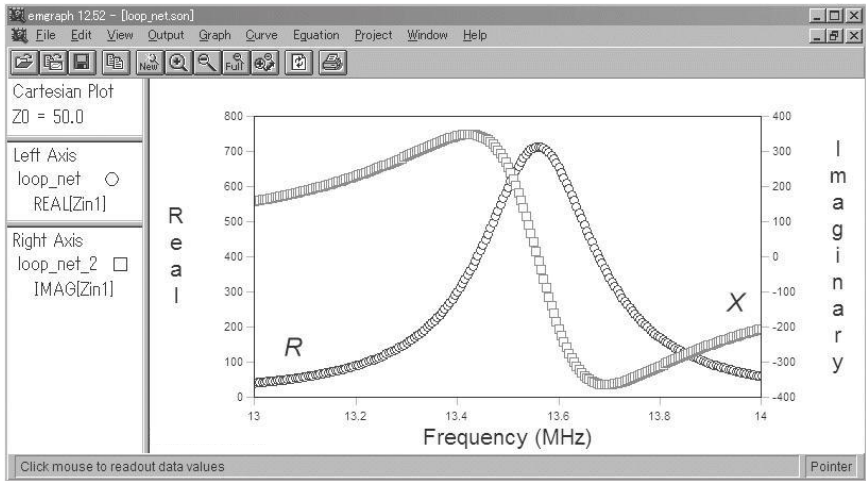


Figure 2.38 Input impedance of this coil is more than 700Ω pure real at 13.56 MHz.

The plot in Figure 2.38 shows input impedance, and reactance is zero at 13.56 MHz. It is a parallel LC resonant circuit and R is more than 700Ω . Typical ICs for this kind of tag are designed for high impedance, so just adding the capacitor finishes this design.

3

Wire Antennas

3.1 Fundamentals of a Dipole Antenna

We have already seen several examples of the dipole antenna. This antenna is such an important building block that we will look at it in even more detail.

3.1.1 Standing Wave on a Linear Dipole Antenna

The dipole antenna (Figure 3.1) realizes desirable performance when its length is about one half-wavelength at the desired frequency of operation. The electric current fed to the center of the dipole launches two traveling waves, one going each way, out toward each end of the dipole. When they reach the end of the wire, they each totally reflect and travel back to the center. If we work the mathematics, a traveling wave going in each direction (out and back) results in a standing wave. Look at the current distribution of Figure 3.1. It is one half-wavelength of a wave that is just standing there. Thus, the dipole is described as a standing wave antenna.

This is illustrated in Figure 3.2. We see two traveling waves (on the left, dashed lines) adding together as they travel along the wire in opposite directions. In 1 through 12, each traveling wave moves an additional $1/12$ wavelength, each moving in opposite directions. The solid line shows the sum of both traveling waves. The sum is also shown all by itself on the right side. At 1, the traveling waves are exactly opposite of each other and they sum to zero when combined. As for 2, the two traveling waves no longer cancel perfectly, and the standing wave increases slightly. From this, we can see that a dipole is

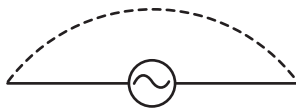


Figure 3.1 Current distribution on a dipole antenna.

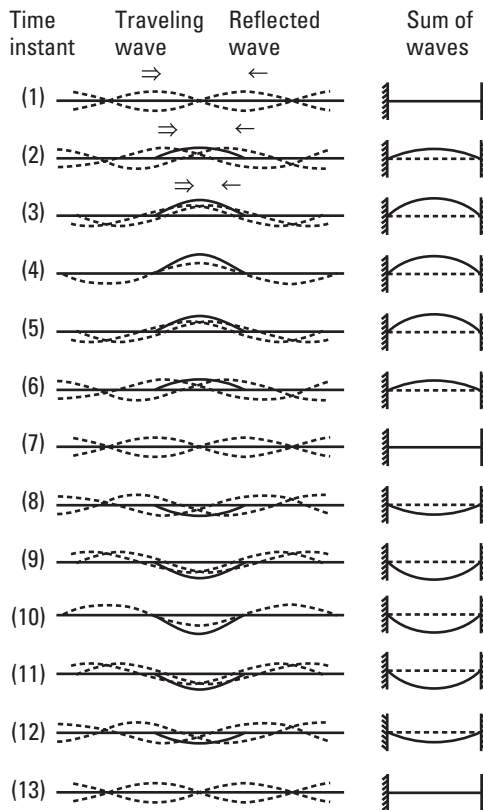


Figure 3.2 Combining two traveling waves that are traveling in opposite directions generates a standing wave.

just like a vibrating guitar string. Both ends of the string are fixed in position and the rest of the string is vibrating.

Because both ends of the dipole antenna are open circuits (i.e., zero current), a standing wave is generated as a sum of the initial traveling wave and the traveling wave reflected from the ends of the wire. This gives us a standing wave, just like on a guitar string.

3.1.2 Designing the Element Length

In textbooks, the input impedance $R + jX$ of a one-half wavelength dipole antenna in free space is $73 + j43\Omega$. The 43Ω reactance means that some of the power that is sent to the antenna is reflected back towards the transmitter, after spending a short time in the complicated fields near the antenna (like energy in the spring of a shock absorber). Reflected power is not radiated. A positive X is inductive reactance and corresponds to an inductor (a coil of wire), which means that extra magnetic field energy is generated. As we have seen in previous chapters, inductance does not need an actual coil of wire. Just a straight wire will do the trick. It turns out that if we cut off about 3% of the dipole wire length, the magnetic field decreases by just enough to remove $+j43\Omega$, and all that remains is the pure resistance of 73Ω . For a thin half-wave dipole in free space, a length of about 97% of one-half wavelength removes the reactance.

The textbook value of $73 + j43\Omega$ is calculated for an infinitely thin, round wire. However we usually build antennas using thick wires and tubes that have a real diameter. Planar antennas use flat “wire” and are placed on a substrate. How do we design such antennas?

3.1.3 A Dipole Antenna on a Substrate

Figure 3.3 shows a linear (straight) dipole antenna. To set up this antenna, select Circuit->Units..., and set cm and MHz. Under Circuit->Box, set the cell size to 0.2 by 0.2 cm and the Box size to 80 by 80 cm. Because this is an antenna simulation, set the Top and Bottom covers to Free Space.

Set Dielectric Layers as shown in Figure 3.4, and draw an element on the dielectric whose thickness is 0.05 cm and relative permittivity is 3. Draw the

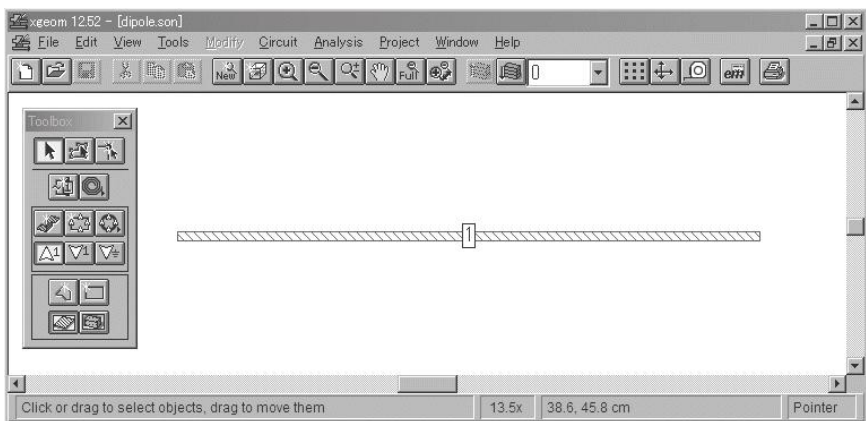


Figure 3.3 A linear (straight) dipole antenna. File: dipole.son

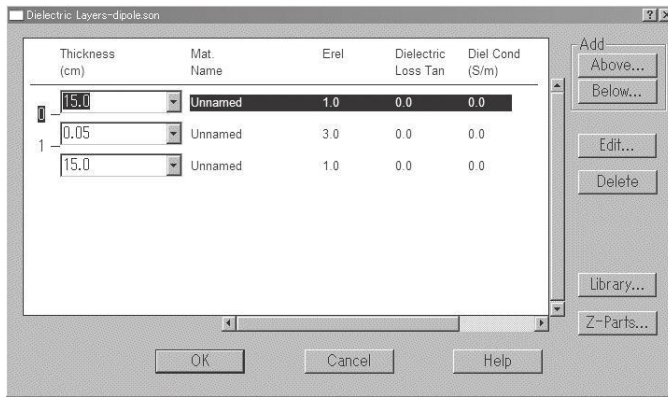


Figure 3.4 Dielectric Layers setup for the dipole on a substrate.

dipole element in two identical pieces, 6.8 by 0.2 cm. Put port 1 at the junction in the center, as we have done for previous dipoles.

3.1.4 Parameterization

Sonnet Lite allows variable parameters and can change them automatically for a simulation. Here we assign the length of the dipole antenna as parameter and so we can design an antenna that resonates at 953 MHz, the operating frequency of UHF RFID (ultrahigh-frequency radio frequency identification) tags in Japan.

After selecting Tools->Add Dimension Parameter->Add Symmetric, click on the upper left corner vertex of the dipole with left button of mouse and that point is highlighted (Figure 3.5). This point is the first anchor point. To set a parameter, follow the next steps. Notice that hints for what to do also appear in the lower left corner of the Sonnet window.

1. We want to select all the points that will also be moved when the anchor point is moved. In this case, there is only one more point. Click on the other vertex for that end of the dipole and press Enter.
2. Now we set the second anchor point. Left-click on the upper far right corner vertex of the dipole and it is selected.
3. Finally, we select all other points that we want to move when this second anchor point is moved. As before, there is only one additional point to be moved. Left-click on the other vertex at this end of the dipole and press Enter.

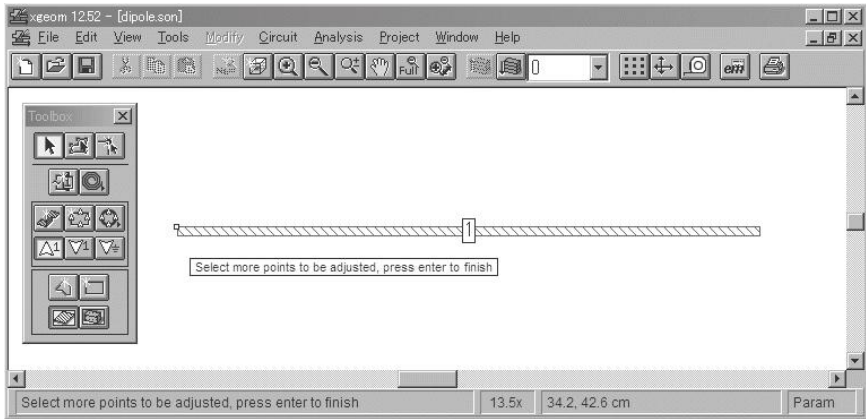


Figure 3.5 Setting an anchor point for the parameterization.

Now the dialog box in Figure 3.6 is displayed. Enter an appropriate variable name, like “length.”

Click the OK button, then a dimensioning arrow is displayed for the parameterized line length (Figure 3.7). Move it to a convenient location. Double-click on the label for full details. With the full details displayed, you can change the parameterized dipole length just by typing in a new length. Try it!

We added a symmetric parameter above. When changing a symmetric parameter, both anchor points are moved an equal distance in opposite directions. Sonnet Lite also has Anchored Parameters (one anchor point moves, the other remains stationary) and Radial Parameters (to parameterize curves).

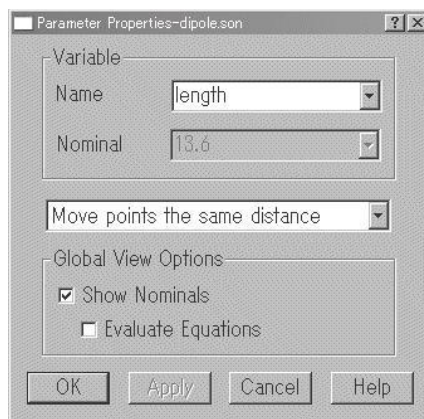


Figure 3.6 Parameter name and other settings.

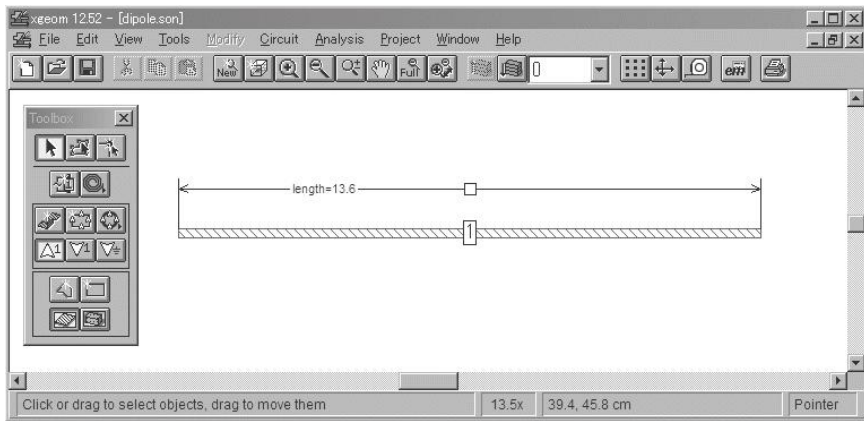


Figure 3.7 A dimensioning arrow labels the parameterized line length.

To set up a parametric analysis, select Analysis->Setup... and set the Analysis Control dropdown menu to Parameter Sweep. Click Add... and the dialog box shown in Figure 3.8 is displayed.

Select an ABS sweep from 800 to 1,100 MHz. Check the length parameter and set it to vary from 12 to 15 cm with a step of 0.2 cm and click the OK button. Then select Project->Analyze.

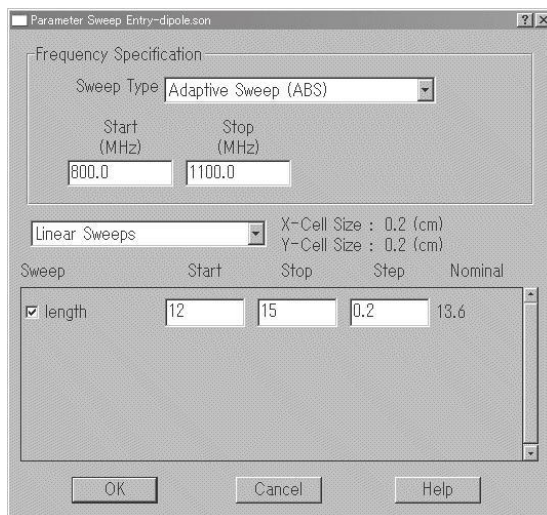


Figure 3.8 Settings for the Parameter Sweep.

When the analysis is complete, select Project->View Response. In the graphing program, double click the “dipole” curve label in the frame on the left-hand side. The dialog box in Figure 3.9 is displayed. Check the Graph All Iterations box and click OK. Figure 3.10 shows the S_{11} magnitude (reflection coefficient) results for all element lengths.

We want the resonant frequency of this dipole to be 953 MHz, and we can check that length = 13.6 cm. To see the length, move your cursor over the desired curve and the value of the length parameter is displayed.

3.1.5 Examining the Impedance

With the plot of Figure 3.10 displayed, select Curve->Add Curves... and change the plotted quantity to the Real Part of Z_{in} . Figure 3.11 shows the result.

In textbooks, the real part of the input impedance of a one-half wavelength dipole is 73Ω . However this planar dipole, which uses flat “wire” and is on a substrate, has an input impedance of 54Ω at resonance, with a dipole length of 13.6 cm (Figure 3.12).

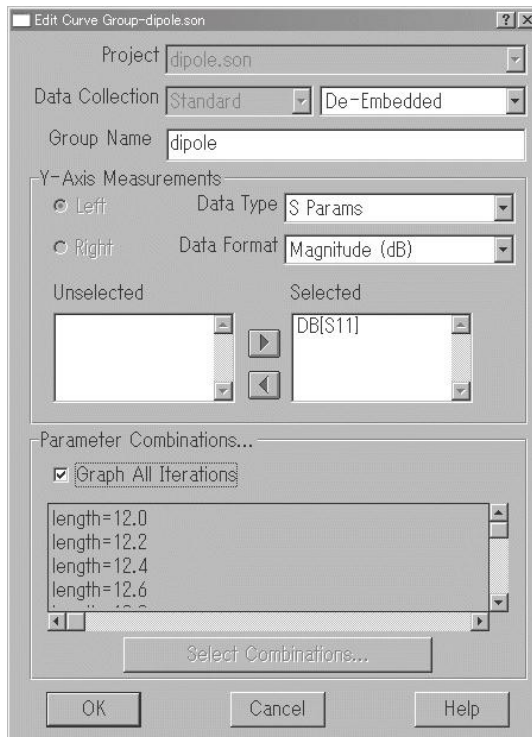


Figure 3.9 Select Display All Iterations.

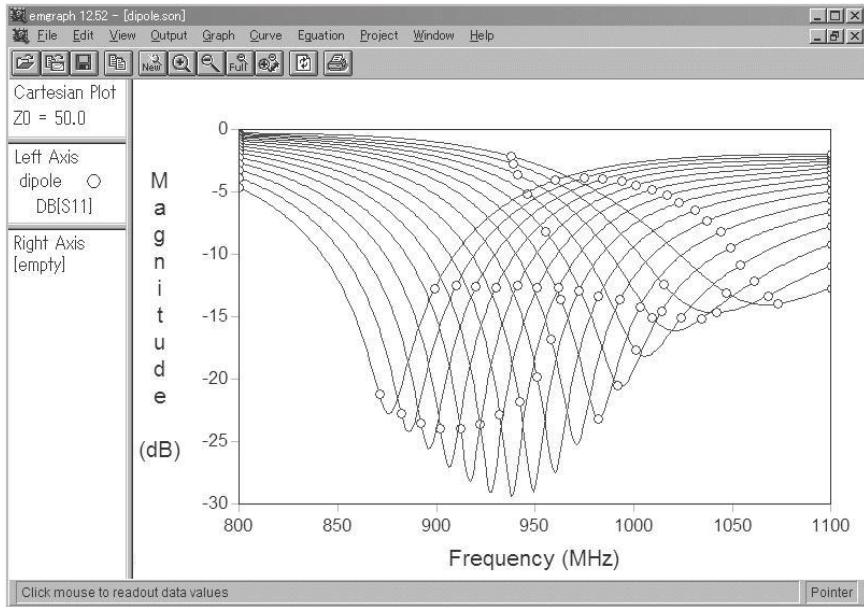


Figure 3.10 Graph of all S_{11} for all the dipoles analyzed.

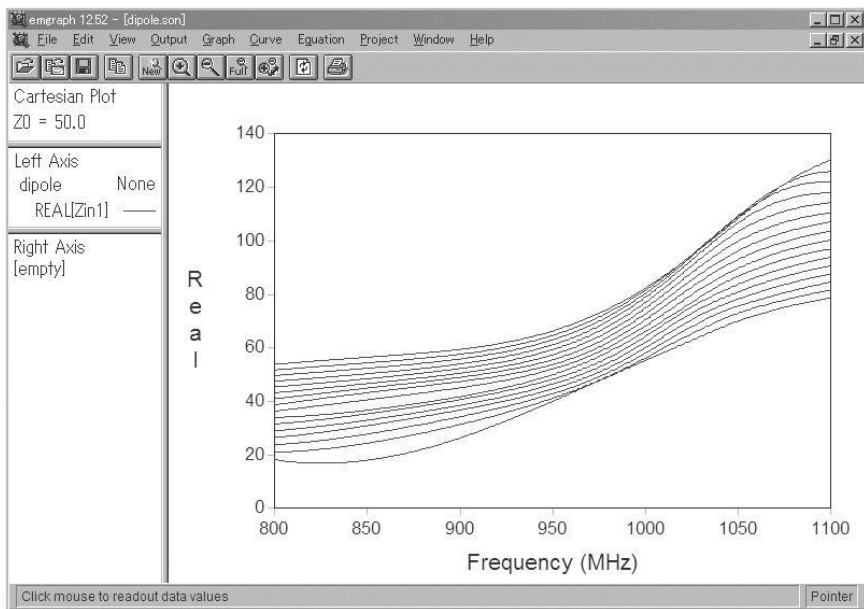


Figure 3.11 Resistance (real part of input impedance) for all element lengths.

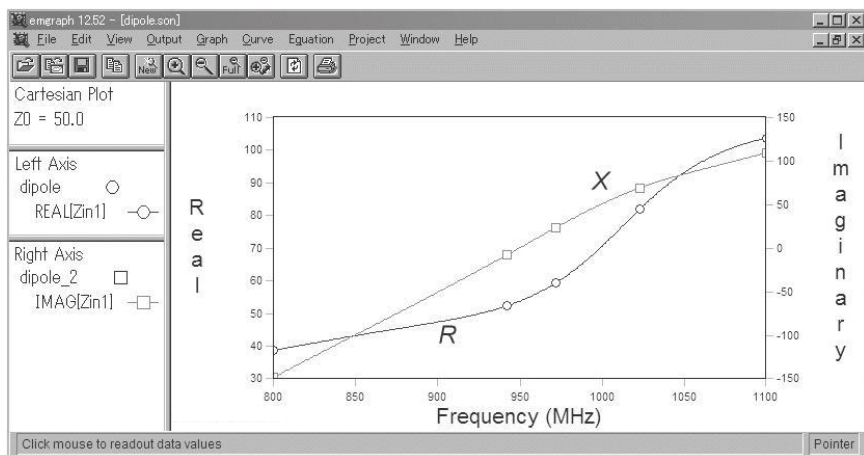


Figure 3.12 Input resistance and reactance for a planar dipole is 54Ω and it is resonant (zero reactance) at about 950 MHz with a length of 13.6 cm.

3.2 Fundamentals of a Loop Antenna

A dipole antenna is typically one-half a wavelength long. A loop antenna is typically one wavelength long. Let's explore the design of a square planar loop antenna.

3.2.1 Simulation for a Quad Antenna

At DC, the loop antenna is a short circuit across the antenna input terminals. At low frequency, in other words, where the length of the antenna is a tiny portion of a wavelength, it is an inductor. It is not really an antenna. Here, we consider a loop antenna whose length is about one wavelength at the desired frequency of operation.

An antenna that is designed so that its loop length is one wavelength and is square is called a quad antenna, which is a familiar term to radio amateurs. Building on the loop antenna of Hertz's receiver from Chapter 2, Figure 3.13 shows maximum current at the feed point (port 1), and there is also a current maximum on the exact opposite side.

When drawing the direction of current with an arrow, the upper half and lower half of the quad are each just like two bent dipole antennas. What is important is the direction of the current. The top and the bottom currents flow in the same direction. Each half radiates. Since the currents of the two halves are in phase, their radiated waves add in phase in the direction perpendicular to the plane of the loop, the direction from which you are viewing the loop. Thus, the antenna pattern of the loop is maximum in this direction.

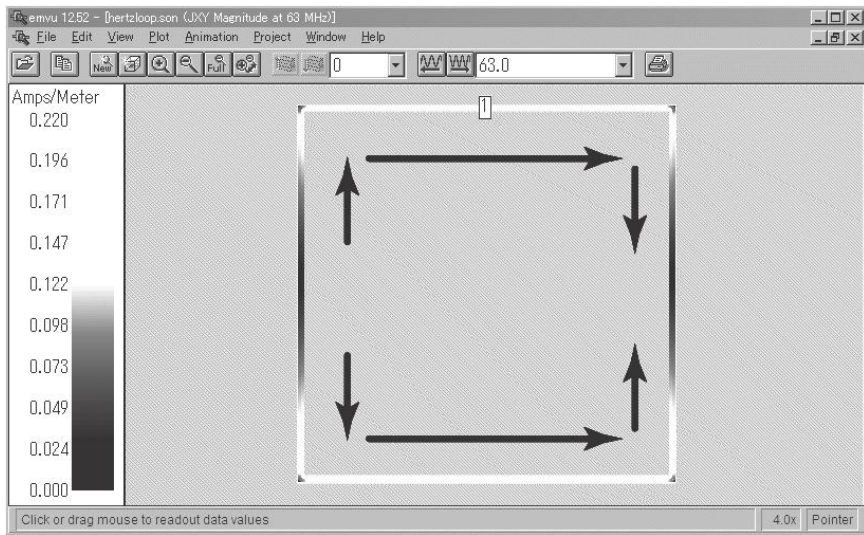


Figure 3.13 Current directions on a quad antenna.

In contrast, in the vertical legs of the loop, the current in opposite legs is oppositely directed. Thus, the radiation from each vertical leg cancels the radiation from the other vertical leg, at least in the direction perpendicular to the plane of the loop.

The input impedance of a quad antenna is shown in Figure 3.14.

At the resonant frequency of 67 MHz, R is 157Ω , about double the dipole antenna. At 125 MHz, where the quad loop is about two wavelengths long, R is 270Ω .

In Figure 3.14, there is another frequency, 89 MHz, where X (reactance) becomes zero. R is as much as 1360Ω , high impedance. This is an antiresonance, or a parallel resonance where the loop is 1.5 wavelengths long (Figure 3.15). The feed point has high resistance because it is at a current minimum, which automatically corresponds to a voltage maximum. Be sure to use a good high-voltage connector at this point!

3.2.2 What Is a Magnetic Loop Antenna?

The antenna for a 13.56 MHz RFID card, like the SONY FeliCa, is a small loop antenna because it is much smaller than the wavelength of 22m. It is a magnetic field type antenna, and the magnetic intensity in the near field is strong. Might it be possible to generate some far-field radiation, even though it is so small?

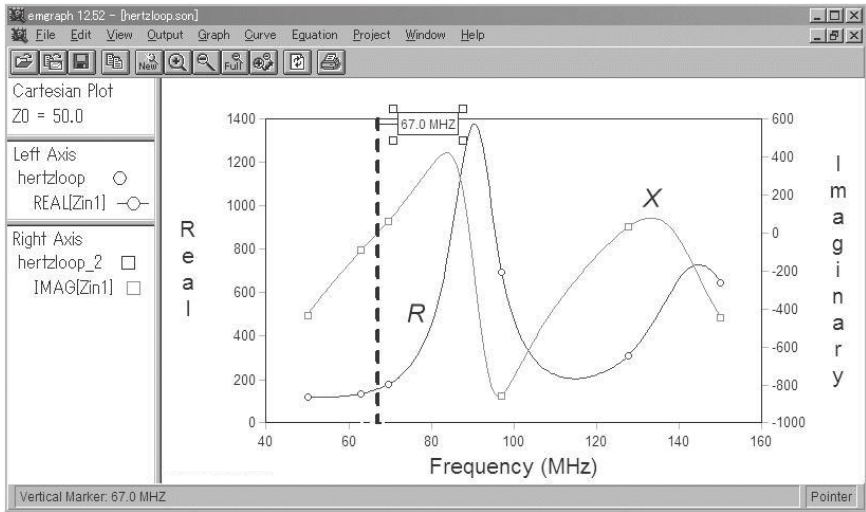


Figure 3.14 Input impedance of a quad antenna.

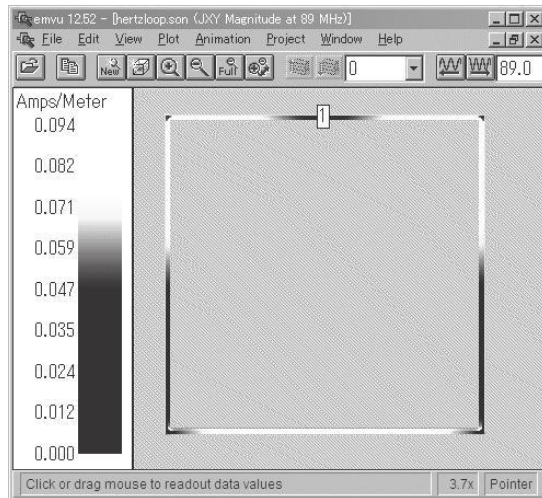


Figure 3.15 The loop is 1.5 wavelengths long at 89 MHz.

In fact, if the loop length is at least one-tenth of wavelength, we can actually build an antenna whose radiation efficiency is more than 50%. This has been a lifelong amateur radio project of one of the authors. We call this type of antenna a magnetic loop antenna (MLA).

Since around 1957 this type of small loop antenna has been studied, and it is based on an antenna called the Army Loop. It was used in the army in the Vietnam War (1960–1975).

Figure 3.16 shows the loop by Patterson (*Electronics*, Aug. 21, 1967, pp. 111–114); this kind of antenna has been investigated for some time. It is important to decrease both the resistive loss of loop conductor and the connector feed point loss. Chris Käferlein (amateur radio call sign DK5CZ) developed such an antenna (Figure 3.17) and has had it in production since 1983. His design was analyzed by Hans Würtz (amateur radio station DL2FA). He has produced many types of this antenna that cover the 160m to 10m bands, which have become popular in Europe. The small coupling loop near the feed is designed for broadband operation with a coaxial feed line.

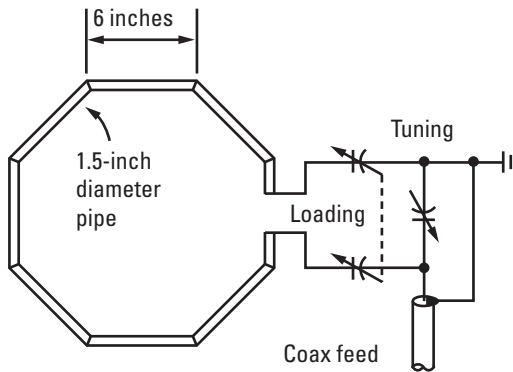


Figure 3.16 Small loop by Patterson (1967).

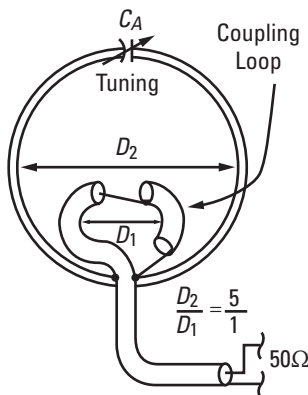


Figure 3.17 Schematic of an MLA by Chris Käferlein.

Figure 3.18 shows one of Käferlein's products, AMA-10D™. The main loop is 1.3m in diameter with a small one-turn loop for matching the coax to the antenna at the bottom. A large variable capacitor is inside the cylindrical case at the top. A remotely controlled motor adjusts the capacitance to change the antenna's resonant frequency.

3.2.3 Simulation of the Small Loop Antenna

Figure 3.19 shows a 100-mm square loop model based on the loop antenna loop.son that we simulated in Chapter 2.

The port is located slightly above the center line; this actually makes little difference. Running a simulation the same way we did for the small loop in Section 2.4.4, the value of L at 13.56 MHz is found to be 371 nH.

Knowing the value of L for this coil, we find that the value C to resonate at 13.56 MHz is 371 pF.

In the example in Section 2.4.4, we designed a parallel resonant circuit by creating a circuit simulator Netlist Project. In this case, we assume that this is a reader/writer antenna for an RFID tag that was designed to operate with a series LC resonant circuit.

We see a small capacitor near port 1 in Figure 3.20. This is an ideal capacitor inserted as a "component." To add a component, cut out a small section of the loop. Do this by selecting Edit->Divide Polygons. Use the little razor knife that appears to cut the loop just below port 1. Select the razor tool again and make a second cut just below the first one. Select the little piece of the loop by left-clicking in it and press the Delete key. This opens up a space for the ideal capacitor.

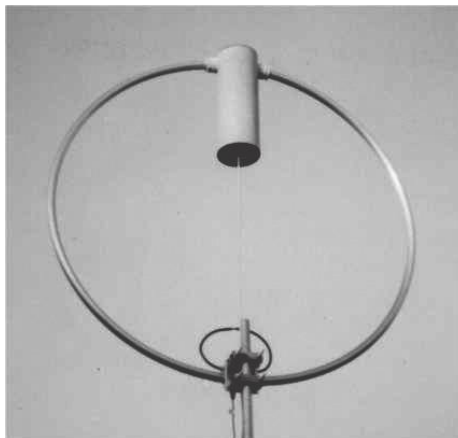


Figure 3.18 MLATM product by Chris Käferlein.

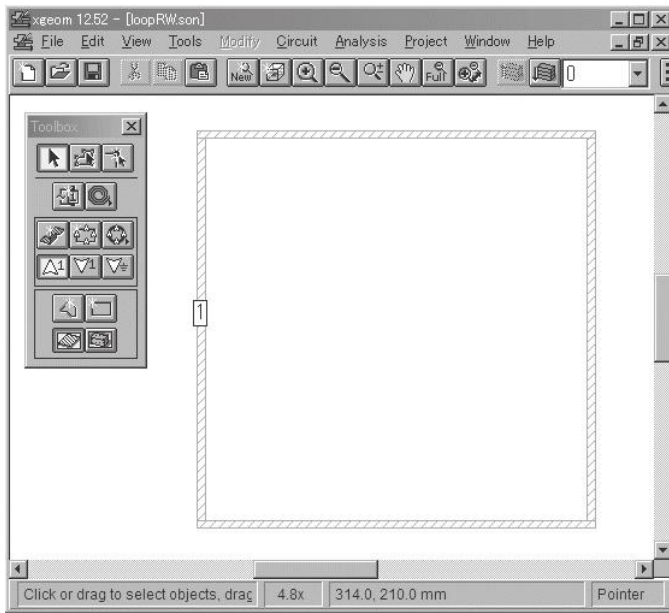


Figure 3.19 Square loop model. File: loopRW.son.

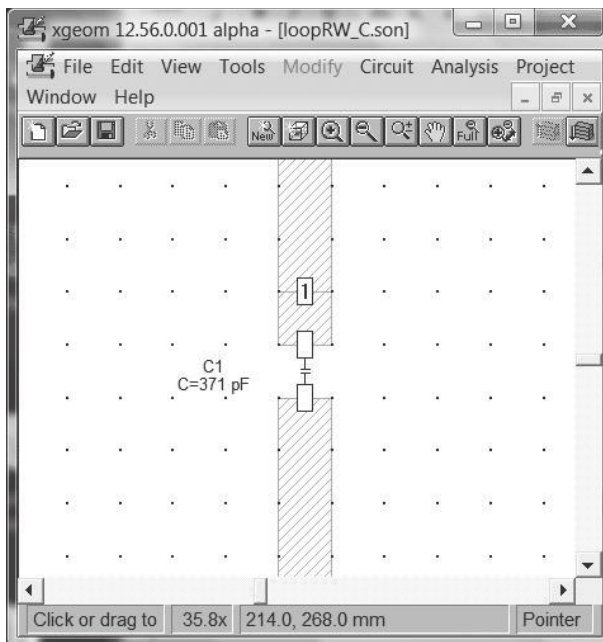


Figure 3.20 Loop with a lumped capacitor included. File: LoopRW_C.son.

Next, select Tools->Add Component->Ideal. The dialog box shown in Figure 3.21 is displayed. Set the Ideal Type drop-down menu to Capacitor. Type in a value of 371 pF. Then, on the right side, set the Terminal Width drop-down menu to Feedline Width.

Click OK and then, following the hints that Sonnet Lite provides, place the two capacitor terminals. After the capacitor is in place, you might want to move the label to a convenient location. Just zoom enough on it to see it clearly, then click on it and drag it to a convenient location. Analyze the circuit and plot the real and imaginary input impedance (Figure 3.22).

Figure 3.22 is the resulting input impedance, showing that R at 13.56 MHz is 0.54Ω and X is nearly zero. The resonant frequency is slightly higher.

Next, in the dialog box displayed after selecting Graph->Terminations..., change the value of Resistance to 0.54Ω and plot S_{11} magnitude dB. Figure 3.23 suggests that it resonates at around 13.645 MHz

Components in Sonnet use internal cocalibrated ports. These are perfectly calibrated internal ports. As such they remove all the inductance associated with the small piece of the loop that was removed. Decreased inductance increases the resonant frequency.

3.2.4 Matching the Small Loop to 50Ω

Figure 3.17 shows the technique used by Hans Würtz to make the antenna feed point impedance 50Ω , using 50Ω coaxial cable. We can imagine that the two coils are the primary and secondary of a transformer. The transformer on the electric power poles outside our homes transforms the very high voltage outside

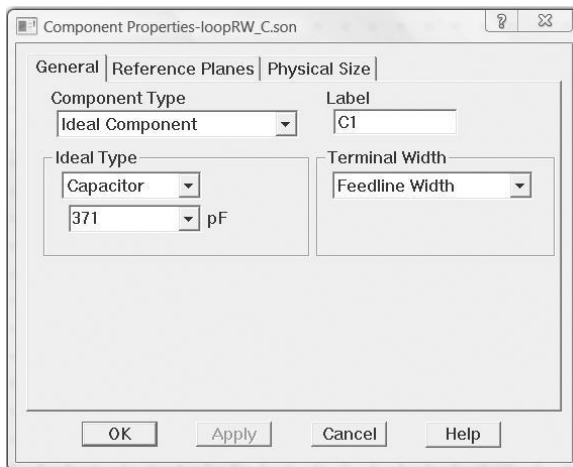


Figure 3.21 The Add Component dialog box.

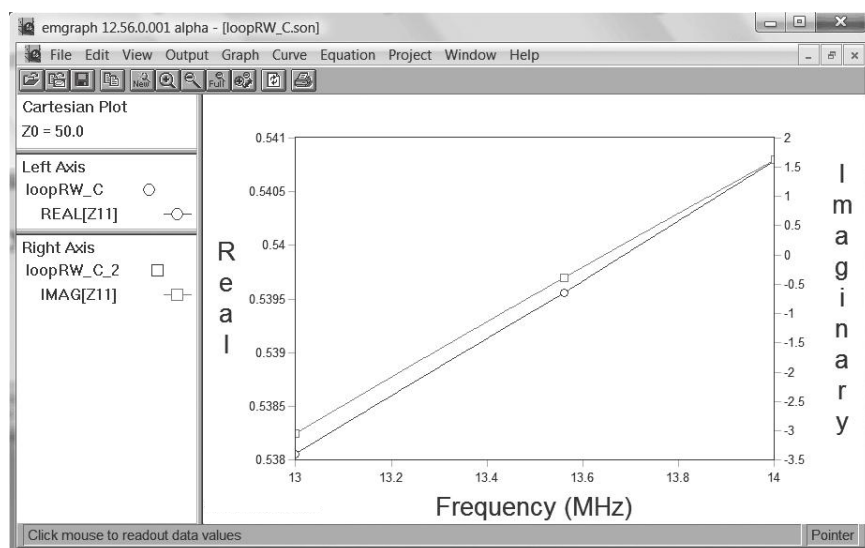


Figure 3.22 The complex input impedance of the loop including the capacitor used to tune the resonant frequency.

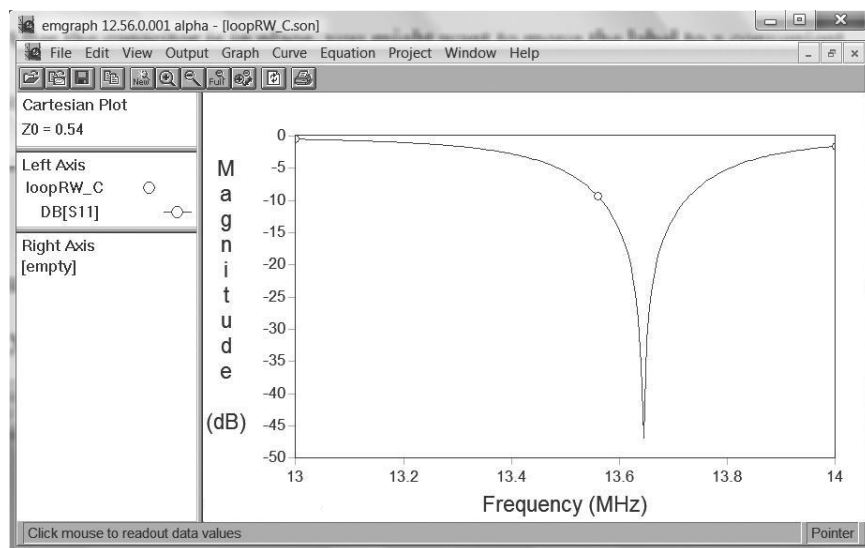


Figure 3.23 The series LC resonance of the RFID loop and tuning capacitor.

down to household voltage levels. It turns out that they also transform impedance. In this way, we can match the small loop to 50Ω .

Figure 3.24 is a model using this idea. Is it possible to adjust the size of the loop to match the loop to 50Ω ?

With the outer coil remaining as in Figure 3.20, delete the port. Next add a loop on the inside that is 60 mm on a side and insert port 1 as shown in Figure 3.24. (Ports can only be added on polygon edges.) Try analyzing a few different sizes. If you like, parameterize the inner loop and run a range of different sizes. Using a loop 60 mm on a side, the input resistance is very close to 50Ω (Figure 3.25).

3.3 Fundamentals of the Yagi-Uda Antenna

From dipoles and loops, we now go to the well known directional antenna, the Yagi-Uda, or as it is more commonly known, the Yagi antenna.

3.3.1 Fundamentals of a Reflector

In Chapter 1, we described a Yagi-Uda antenna with two elements. Here we examine fundamentals of the Yagi-Uda antenna based on the book by Dr. Shin-ichiro Uda, one of the inventors.

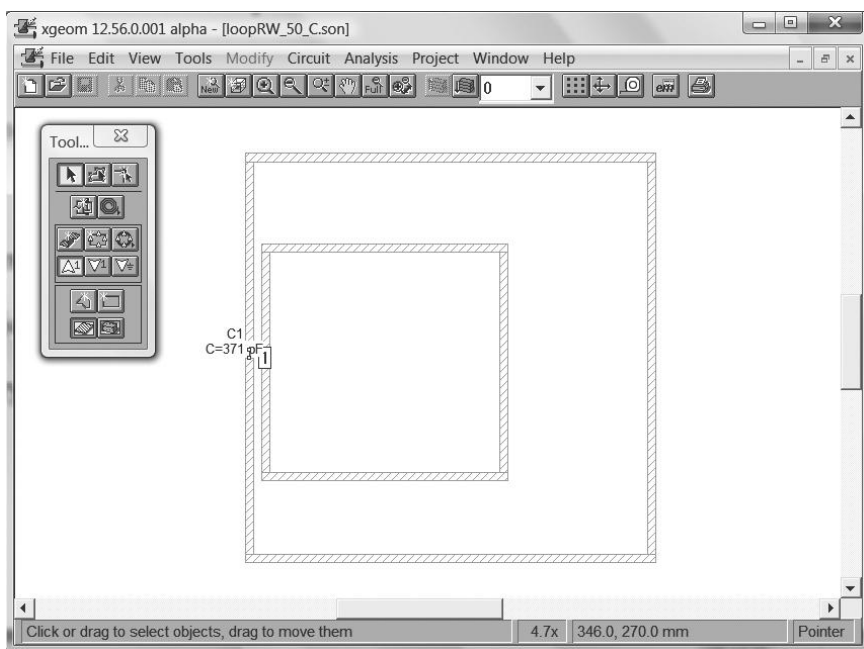


Figure 3.24 Impedance matching using a small inner loop. File: loopRW50.son.

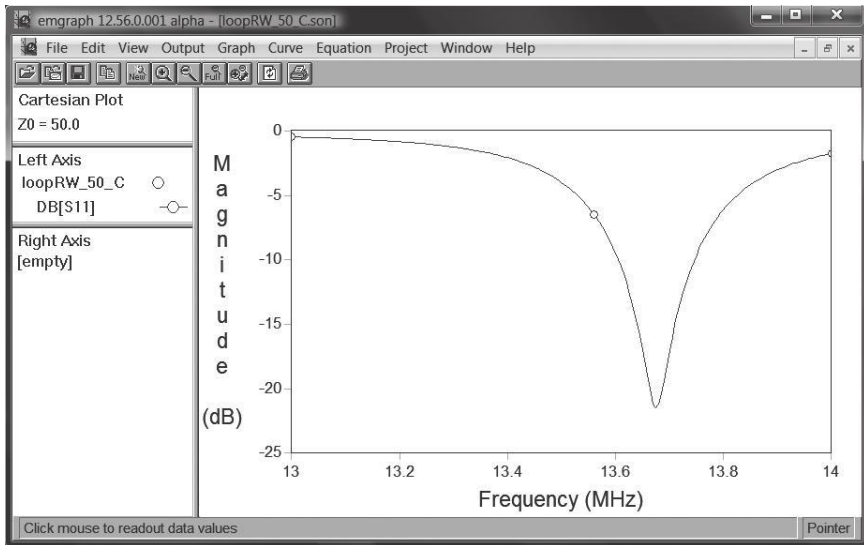


Figure 3.25 The small impedance matching loop almost perfectly matches the loop to 50Ω at 13.645 MHz, S_{11} in a 50Ω system is plotted.

The element A in Figure 3.26 is a side view of a horizontal $\lambda/2$ dipole antenna. R is also a dipole antenna, whose length is slightly longer than $\lambda/2$. The distance between the two elements is $\lambda/4$.

For a Yagi antenna, the feed line is connected only to what is called the “driven element,” element A. When feeding element A, current on element A induces current on the nearby element R even though the feed line is not connected to element R. It does not matter whether the current on an element comes from a direct feed line connection, or from being induced by current flowing on a nearby element. All current on all elements, both A and R, radiate. Radiation from the director, A, is shown with a solid line. Radiation from the reflector, R, is shown with a dashed line. In both cases, traveling waves are launched by the elements, and the traveling waves travel, of course, away from

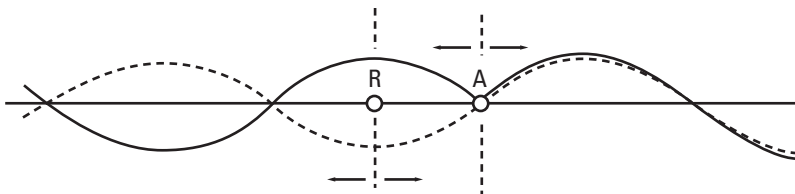


Figure 3.26 Conceptual side view of a Yagi antenna. Waves to the right are in phase and to the left are reversed phase.

the radiating element. The waves traveling to the right are in phase and reinforce. At the same time the waves to the left are reversed in phase and cancel each other.

Because the distance between A and R is $\lambda/4$, the excitation of R by current on A is delayed by approximately 90 degrees. R is made a bit longer than $\lambda/2$ so that the current in R is delayed by about 180 degrees. Because of this, the wave reradiated from R is now delayed a total of 270 degrees with respect to the radiation from A.

Next, the wave radiated from R must travel the 90 degree delay from R back to A again, adding another 90 degrees of delay, the total delay is now 360 degrees, which is electrically the same as 0 degrees, and the waves from both A and R are in phase going to the right. Consequently, the solid line traveling from A and the dotted line reinforce each other to the right, Figure 3.26.

And as for the waves radiated to the rear (to the left), the solid line and the dotted line are of opposite phase, and they cancel. This is called the reverse direction, or “off the back of the antenna.” However as described in Chapter 2, the near field of a dipole antenna is not simple. So we need to view Figure 3.26 a conceptual explanation. In actual design there is a complicated trade-off between making the reflector longer and where it is positioned in back of the director. A lot of time can be spent optimizing a Yagi for maximum performance.

3.3.2 Fundamentals of a Director

If making an element a little longer than $\lambda/2$ makes it into a reflector, perhaps making it a little shorter changes it into a director. A director, D, is placed $\lambda/4$ in front of A, and its length is made shorter so that the current in the director is advanced by 90 degrees. This means that the induced current in the director generates a traveling wave that reinforces with the traveling wave from the driven element, and we get even stronger radiation in the forward direction. Of course, this also means, by conservation of energy, that the total radiation elsewhere (in this case, to the rear and sides of the antenna) is decreased. Thus, the Yagi can be a highly directional antenna.

The impedance of D is influenced by A, and A is also influenced by D. The same is true for the reflector, R. Because the interactions are so complicated, an actual Yagi-Uda antenna must be optimized by trying different element lengths and spacings. While there is typically only one reflector (at least when all the elements are on a single boom), there can be many directors. Additional directors increase the forward gain...if they are of the right length and spacing. Finding an optimal Yagi for a given number of directors on a given boom length is a difficult problem. In the past it has been done by physically building different Yagi antennas and measuring them. The best one is selected. Now this is done using computers.

The Yagi-Uda antenna was invented in the end of Taisho Era, and it was in 1928 that their work was presented in English. The next year a UHF (at 670 MHz) Yagi-Uda antenna successfully established communication over a path of 20 km in Sendai. It is on display in the Research Institute of Electrical Communication, Tohoku University (Figure 3.27 is offered by Mr. Hironobu Hongou, Fujitsu Limited).

3.3.3 Simulation of the Yagi-Uda Antenna

Sonnet Lite provides for two dimensional drawing, so it is most suitable for planar antennas on substrates, Chapter 4. Wire antennas can be designed with another free software package. This is an antenna analysis named MMANA written by radio amateur JE3HHT, Mr. Makoto Mori, based on the method of moments. The method of moments is a technique that divides the metal of a circuit (as with Sonnet) or antenna (as with MMANA) into a fine mesh of tiny subsections and transforms the problem into matrix equations (simultaneous linear equations) for numerical solution. It can analyze antennas composed of wires and pipes. MMANA, can be downloaded from <http://www33.ocn.ne.jp/~je3hht/> for the Japanese version and the multilingual version, MMA-NA-GAL, from <http://mmhamsoft.amateur-radio.ca/>.

The effect of ground on radiated waves is determined by the combining direct wave (which goes from the antenna directly up and out into space) and the reflected wave (which is radiated from the antenna down towards the earth, then reflects from the earth back up into space), the radiation patterns differ according to the height and location. For the mathematics, the ground is defined



Figure 3.27 Original Yagi-Uda antenna for UHF (670 MHz) at Tohoku University.

as a conductive dielectric. Different kinds of ground have different effects on the radiation characteristics of antennas.

When starting up MMANA or MMANA-GAL, after selecting File->Open, you can use many example files included in ANT folder in MMANA folder. Figure 3.28 shows the screen displayed after opening 8EL6M.MAA and loading antenna data in memory.

Under the Geometry tab for antenna definition, a coordinate system for the end points of wire elements is displayed in the upper half of the screen.

After clicking the View tab for the antenna structure, the Yagi-Uda antenna with eight elements for 50 MHz is drawn in Figure 3.29. Here, the centrally located $\bigcirc$ mark is a feed point.

The origin of the coordinate system is located at the feed point. This displays the three dimensional coordinate system as described for the Geometry tab.

Figure 3.30 shows the Calculate tab where frequencies and environments are set. The height of the antenna and conductor properties are also defined here.

The antenna environment choices are free space, perfectly conducting ground, and real ground. After selecting real ground, in the dialog box displayed after clicking the Ground setup button to the right, specify the specific relative permittivity, ϵ_r , and conductivity, σ , (milliSiemens/meter). Approximate numbers are:

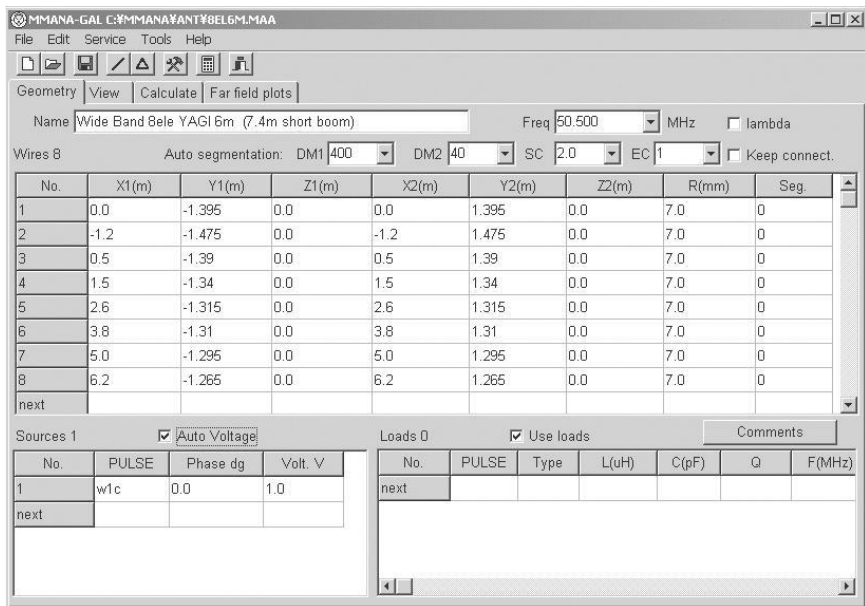


Figure 3.28 MMANA-GAL opens an example file, 8EL6M.MAA.

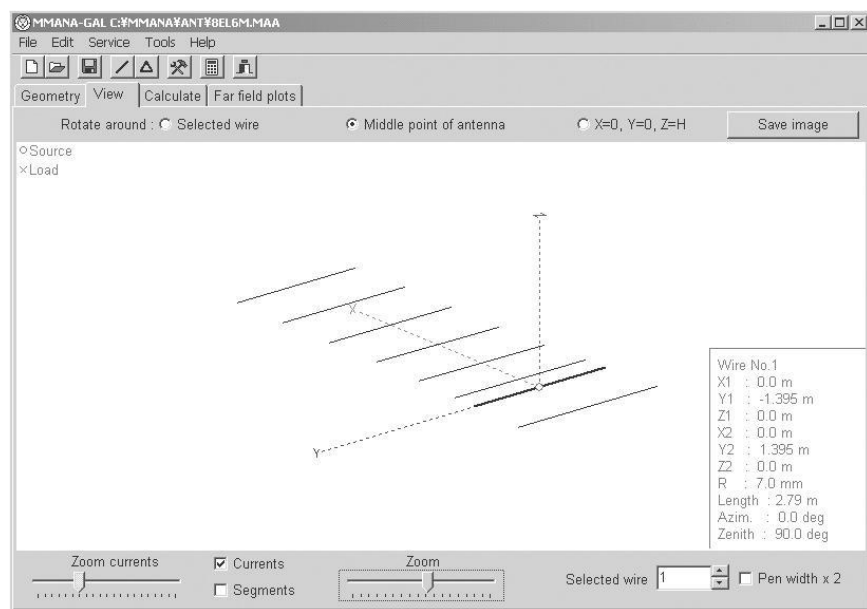


Figure 3.29 50 MHz 8 element Yagi-Uda antenna.

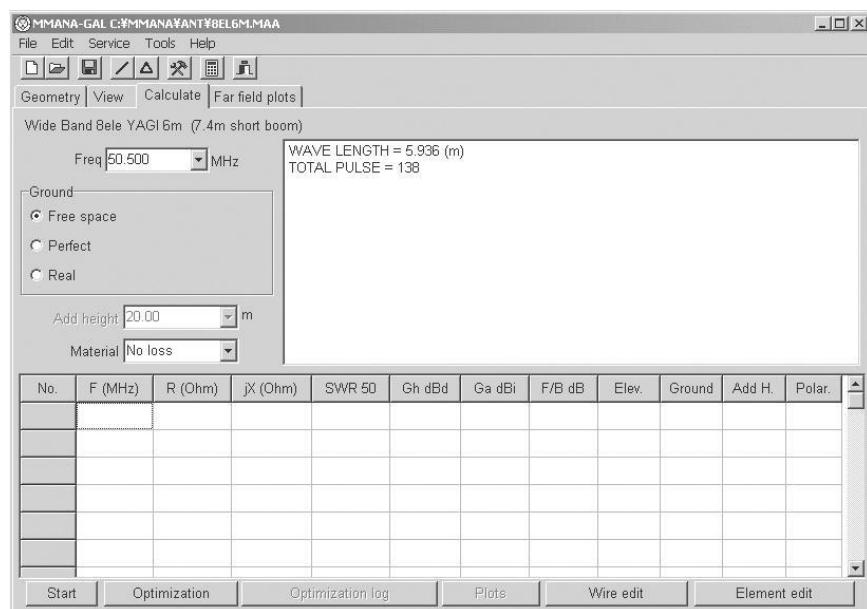


Figure 3.30 Calculation tab where frequencies and environments are set.

Ocean water: $\epsilon_r = 80$, $\sigma = 1000 \sim 5000$ mS/m

Damp ground: $\epsilon_r = 5 \sim 15$, $\sigma = 1 \sim 10$ mS/m

Dry land: $\epsilon_r = 2 \sim 6$, $\sigma = 0.1$ mS/m

After clicking the Start button, lower left in the Calculate tab screen, the result will be displayed as shown in Figure 3.31. In the large frame at the upper right, antenna input impedance and SWR are displayed. It shows No Fatal Error(s), so we see that it finished without any errors.

Next, click the far field plots tab and the radiation pattern is shown in Figure 3.32 appears. The half circle to the right is the radiation pattern on a vertical cross section (x-z plane) seen from the horizontal direction. The information below that is:

First line: Ga: 17.09 dBi = 0 dB (Horizontal polarization) means that the absolute gain compared to a theoretical isotropic antenna is 17.09 dB. It is not possible to build a real isotropic antenna, but is easy to calculate and is often used for comparison. It has the amazing property that it radiates equally well (a gain of 0 dB) in all directions. The plot of horizontal polarization has the outer edge of the plot labeled 0 dB. That means when

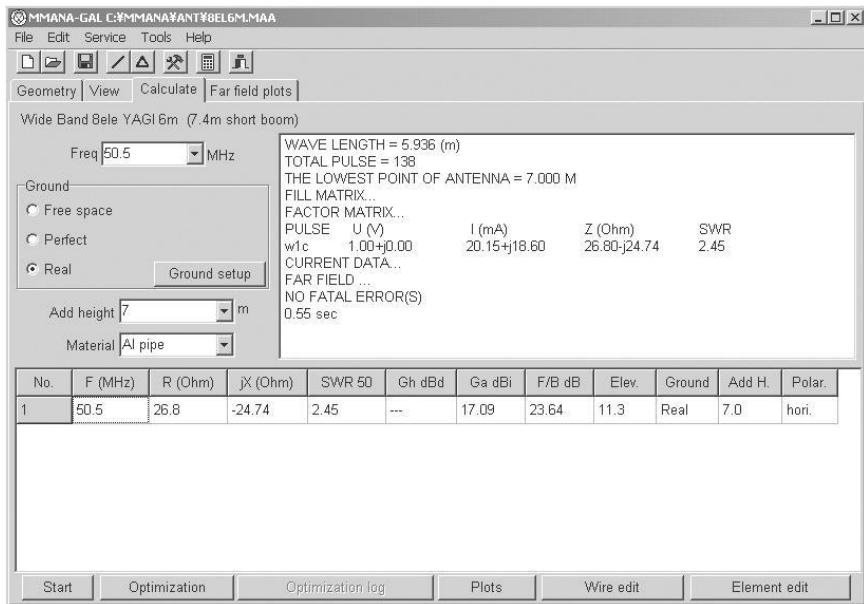


Figure 3.31 Result is displayed in the Calculate tab screen.

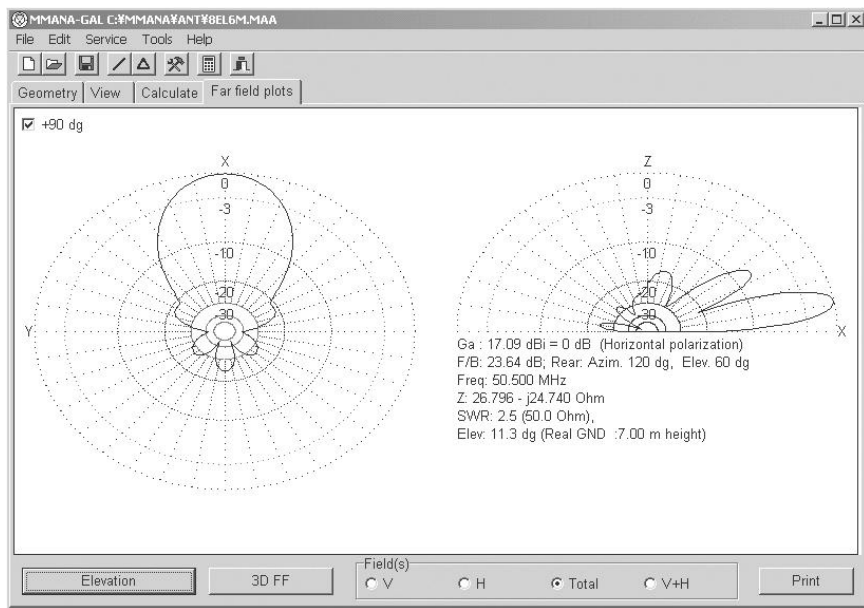


Figure 3.32 Radiation pattern in the Pattern tab screen.

you read the gain from the plot, you should add 17.09 dB to get dBi. The center of the plot is $-\infty$ dB.

Second line: F/B is the front-to-back ratio. This how much stronger radiation is in the forward direction as compared to the reverse direction (off the back of the antenna). The direction of the strongest radiation is defined to be the forward direction.

Fourth line: Z is the antenna input impedance.

Fifth line: SWR is a standing wave ratio.

Sixth line: The strongest radiation is at an angle 11.3 degrees above the horizon, in this case, above a real ground with resistive losses.

First a quick word about SWR. When you have a perfectly matched antenna, the traveling wave launched by the transmitter goes out to the antenna and is completely absorbed either by radiation or by losses in the antenna and matching network (if any). The voltage along the feed line changes in phase, but is always the same magnitude (ignoring loss). If the antenna is not perfectly matched, then a portion of the wave is reflected by the antenna back towards the transmitter. When you have two waves on a feed line, each traveling in

opposite directions, you get a standing wave. When there is more reflected wave you get larger standing waves. The standing wave ratio (SWR) is the ratio of the voltage at the peak of a standing wave to the voltage at the minimum of a standing wave. A perfect match is an SWR of 1.0. If everything is reflected, the SWR is infinity. In practice, an SWR of 2 or less is often considered reasonably good.

The radio buttons displayed in the lower right edge of the window specifies the component of polarized wave that you want displayed. Selections are just vertical, just horizontal, both horizontal and vertical, and the sum of vertical and horizontal.

When clicking the View tab after calculation, the current distribution is displayed, Figure 3.33. This is useful for understanding conceptually what is happening in your antenna.

Electrical current cannot flow beyond the ends of an element, so the current goes to zero at the end, and strongest current typically flows mostly in the center of an element.

When you accustom yourself to using MMANA, you might want to change the settings for meshing (how finely the wire is divided for analysis) and adjust the analysis accuracy. When you check Segments (bottom left), the meshing is displayed on the wire elements with small $\times$ marks.

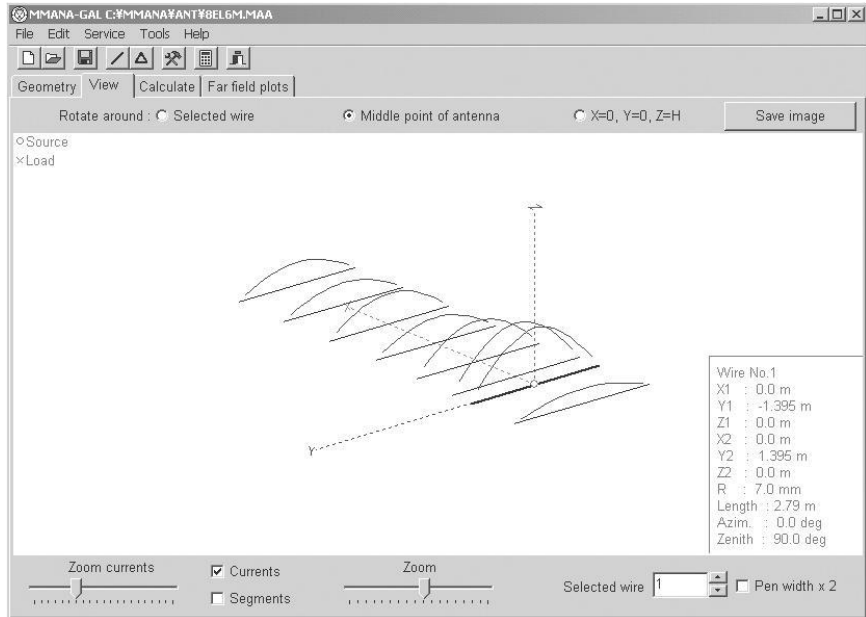


Figure 3.33 Current distribution can be displayed.

3.4 Importance of Antenna Input Impedance

We have touched briefly on antenna input impedance in previous sections. Now it is time to deal with the topic in more detail.

3.4.1 A 50-Ohm Dipole Antenna

Resistance in coaxial cable conductors causes resistive transmission line loss. It turns out that there is a characteristic impedance for coaxial cable that minimizes resistive transmission loss. For example, when coax uses an air dielectric, we see minimum loss in the cable when we choose its dimensions so that the characteristic impedance (Z_0) is about 75 ohms. When the coax is filled with a dielectric of relative permittivity of 2.55 (say, polyethylene), the minimum loss dimensions give a Z_0 of about 50 ohms. And in fact, coax with a Z_0 of 50 ohms is often used.

A dipole has a free space input impedance of 73 ohms. Typically, we want to feed it with 50 ohm coax. At low frequency, dipoles are often relatively close to a real ground, and this can lower the input impedance of a dipole to close to 50 ohms. However, at higher frequencies, dipoles might be many wavelengths from ground. Is there any way we can change a dipole so that it has a convenient input impedance of 50 ohms? Yes, there is. Figure 3.34 shows a dipole with both ends of the antenna bent at 90 degrees.

When bending a dipole as shown, the electric field and magnetic field distributions around the antenna change and the input impedance decreases from 73 ohms. With the right amount of bending, we can realize a free space dipole with 50 ohm input impedance.

The plot in Figure 3.35 shows the antenna impedance with zero reactance (imaginary) and 50 ohm resistance (real) at 953 MHz.

Figure 3.36 shows a plot of the S_{11} magnitude, or reflection coefficient in dB. When we connect a 50 ohm coaxial cable to this antenna, we might think

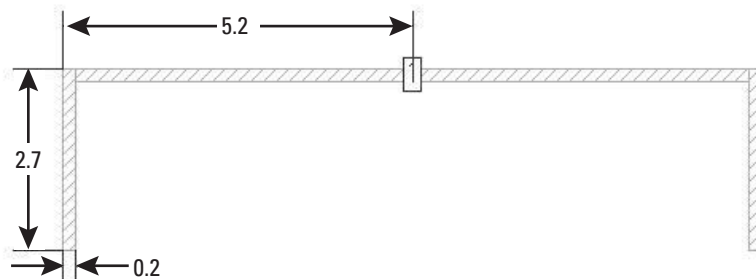


Figure 3.34 Bent dipole model (unit: cm). File: bentdp.son

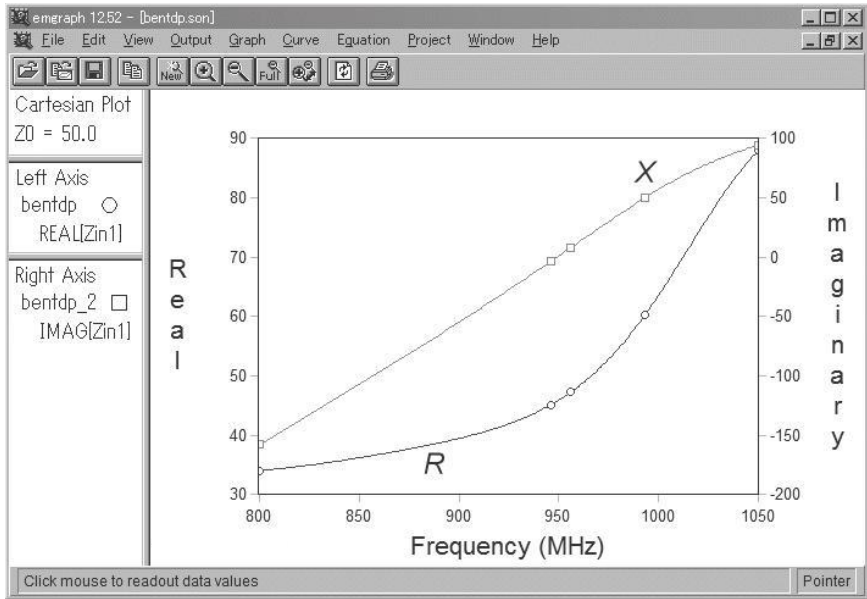


Figure 3.35 Input impedance of a bent dipole.

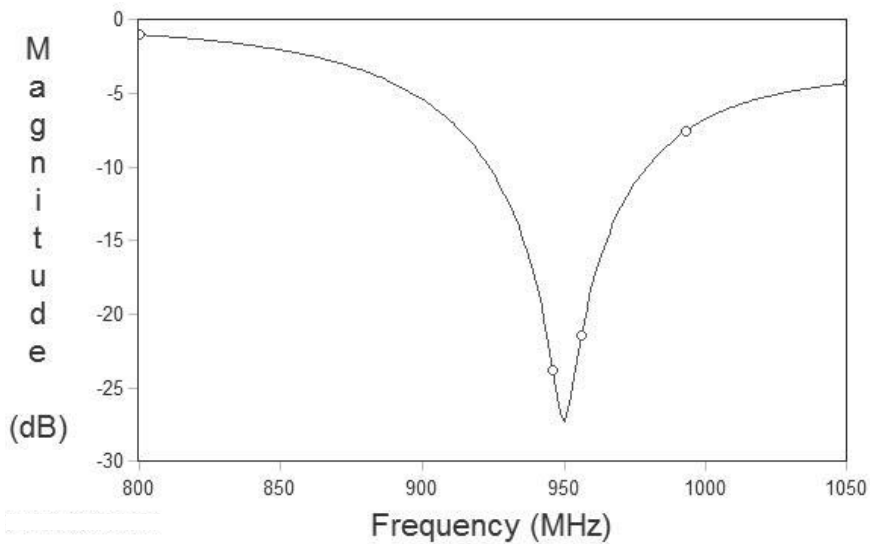


Figure 3.36 Reflection coefficient of a bent dipole.

that there would be no reflection because it is well matched. However directly connecting a coaxial cable might cause another problem.

3.4.2 What is a BALUN?

The dipole antenna is a balanced circuit, in that it is perfectly symmetric about the feed point, right and left. Coax however, is not balanced. It has a center conductor and an outer conductor and they are not symmetric in that you cannot swap one with the other and still have the same thing. When connecting unbalanced coaxial cable directly to a balanced antenna, like Figure 3.37, it is possible for current to flow on the outside of the outer conductor. This current radiates just like any other current and this can adversely affect your radiation pattern. In extreme cases, the current can get all the way back to your transmitter and cause problems there. (Amateur radio operators refer to this as “RF in the shack.”) When receiving, the outside of the coaxial cable acts as a receive antenna, receiving environmental noise. Keep in mind that problems are not guaranteed. Depending on the sensitivity of your systems and the specific orientation between the coax and the antenna, there might be no problem at all.

When directly connecting a balanced line like a ribbon feed line or “twin line” to an unbalanced coaxial cable, Figure 3.38, unwanted radiation from both lines occurs, and the shielding effects of the coax can be lost. In addition, the Z_0 of coax and twin line are usually substantially different, so there can be substantial reflections at the junction.

The direct connection of unbalanced lines and balanced lines (or dipole feed points) can cause problems. The solution is to make the connection using a BALUN, for example, the Sperrtopf BALUN, Figure 3.39, which eliminates current on the outside of the coax.

A well known characteristic of standing waves on transmission lines is that at the high voltage peak of a standing wave, the current is zero. It looks just like

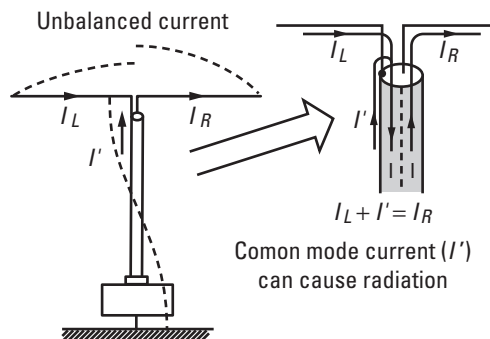


Figure 3.37 Connecting a coaxial cable directly to a dipole antenna might cause problems.

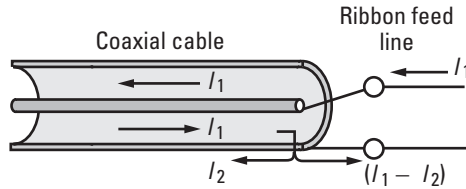


Figure 3.38 A ribbon feed line connected to a coaxial cable.

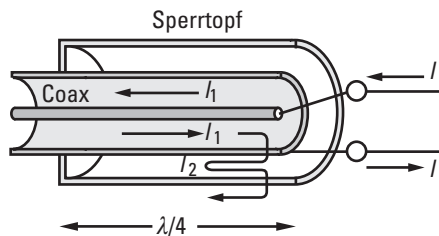


Figure 3.39 Schematic of the Sperrtopf. The large sleeve is shorted to the coax outer conductor at the far left end of the sleeve.

an open circuit. One quarter wavelength away, we have a low voltage null and high current. It looks just like a short circuit. The Sperrtopf takes advantage of this. We want an open circuit for the outer coax conductor where it joins with the twin line. So, we take a tube that is larger in diameter than the outer diameter of the coax and slip it over the end of the coax. We make the tube precisely $\frac{1}{4}$ wavelength long at the operating frequency. Then we short circuit the left end of the sleeve to the outside of the outer shield of the coax. This is a short circuit and creates a standing wave inside the sleeve with the high current (short circuit) point where the sleeve connects to the coax shield. One quarter wavelength to the right, we have an open circuit on the outside of the coax shield, just what we want! Now no current can get onto the outside of the coax shield.

When you calculate the length of a $\frac{1}{4}$ wavelength, be sure to take into account the dielectric constant of the insulation on the outside of your coax. Typically it will be around 2.5 or so. This means that $\frac{1}{4}$ wavelength is about $\frac{2}{3}$ of what you would calculate in free space (one over the square root of 2.55).

A circuit that connects an unbalanced circuit to a balanced circuit is called a BALUN (Balanced to Unbalanced Transducer). There are many forms for a BALUN. The BALUN described above is made for a specific frequency. If you want it to work at a different frequency, you have to change the length. Another form for a BALUN is a transformer wound on a toroidal (i.e., circular) magnetic ferrite core. Like the above balun, it presents an open circuit for any

current trying to flow on the outside of the coax shield. However, this kind of balun tends to be broad band, working over several orders of magnitude of frequencies. In addition, just like with a transformer, you can design the balun to transform between different impedances. For maximum bandwidth, integer ratios of impedance transformation are desirable, like 1:2, or 1:3, etc.

3.4.3 What is the Matched Load?

As we have learned in Chapter 1, an electric circuit is composed of three elements, a power supply, a transmission line, and a load, Figure 3.40. For transmitting an electromagnetic wave, the three elements are a transmitter, a transmission line (like coax), and an antenna.

Under perfect circumstances, an antenna radiates 100% of the electrical power from a transmitter. However, to do that (in typical cases) the transmitter is matched to the characteristic impedance of the transmission line, which should equal the input impedance of the antenna.

In typical high frequency circuits, we usually choose this impedance to be 50 ohms. In some cases (for example, cable TV systems), this impedance is chosen to be 75 ohms. The reason all the impedances in a system should be equal is seen in Figure 3.41.

When the internal resistance of a power supply is R_i and the load resistance is R_L , the condition for maximum power supplied to the load is $R_i = R_L$, shown Figure 3.40. From the viewpoint of the transmitter, R_i is the impedance that the transmitter is designed to drive and R_L is the characteristic impedance of the transmission line. From the viewpoint of the transmission line, R_i is the characteristic impedance of the transmission line and R_L is the input impedance of the antenna. (It is a little more complicated if any of the impedances are complex, we skip that topic to keep things a bit easier.)

When this condition is met, we say that “the transmitter is matched to the load”. When we are using 50 ohm coaxial cable, it is best to design the antenna so that the input impedance is also 50 ohms.

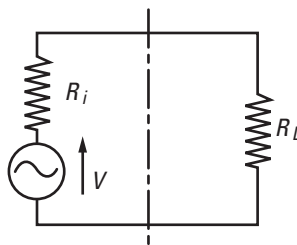


Figure 3.40 The three elements of an electric circuit: the source, the load, and the connection between them.

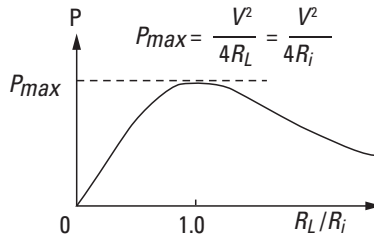


Figure 3.41 Maximum power is transferred from the source to the load when their impedances are all equal.

3.4.4 Need for a Matching Circuit

When miniaturizing a dipole antenna, R might drop to below 73 ohms and capacitive reactance ($-jX$) increases. Modifying the shape of the antenna, like the bent dipole above has its limits. So, in some situations, a matching circuit is needed.

One way this can be done is by using a network of capacitors and inductors. We discuss this in detail in Section 8.5.3.

3.5 Instruments for Measurement of Input Impedance

The complex input impedance of an antenna is critical for proper design and operation. Computer simulation can only go so far. At some point we must build and measure our antennas. This is done with network analyzers. Figure 3.42 shows an example of a vector impedance analyzer and Figure 3.43 shows an example of a vector network analyzer (VNA). The name vector comes from the fact that they can measure the complex impedance of circuits connected to them. A complex impedance has a real and an imaginary part. It can also mathematically be converted to a magnitude and an angle (which is where “vector”

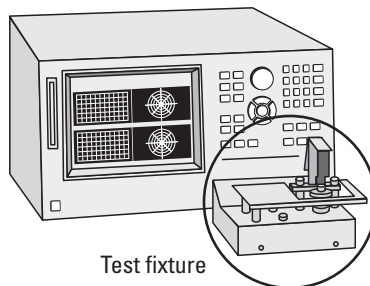


Figure 3.42 Vector impedance analyzer.

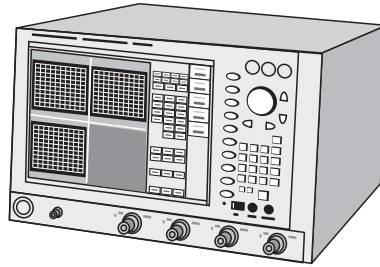


Figure 3.43 Vector network analyzer (VNA).

in the name comes from). Often, we just call them by the general term network analyzer.

In order to get the impedance of a high frequency device, we often use these instruments with special test fixtures (Figure 3.44). When using a VNA, the impedance can be measured after connecting to the feed point of a prototype antenna. For accurate measurements, it is important to calibrate the network analyzer. For example, we typically want the impedance of the antenna. We are not interested in the impedance of the antenna combined with the coax used to connect the antenna to the network analyzer

There are many different VNA calibration techniques. One is the OSL method which uses three standard, well characterized terminations: open, short and load (matched load: 50Ω). The calibration procedure for a given network analyzer is typically contained in a computer program that comes with the network analyzer and runs on a computer inside the network analyzer. You should carefully follow the instructions provided with the network analyzer.



Figure 3.44 Test fixture for a balanced film type antennas (Courtesy Candex Systems, Inc.).

Sometimes, special situations require special calibration procedures. For example, RFID tag elements for UHF (ultra high frequency, which includes anything from 300 to 3,000 MHz) are printed on a PET film or on a paper with a special conductive ink. In order to measure their impedance, we can use a fixture specialized for balanced film type antennas (Figure 3.42: Cadox Systems, Inc.).

4

Antennas on Substrates

4.1 Substrate Dielectrics and Wavelength Shortening

Whenever we bring any dielectric near an antenna, the resonant frequency of the antenna is lowered. Let's explore this phenomenon quantitatively.

4.1.1 Dipole Antenna on a Substrate

Figure 4.1 is a model based on the dipole antenna in Chapter 1 (file name: dp.son). We changed the line width to 2 mm and the element length to 240 mm, which is about 96% of one-half a wavelength at 600 MHz. The reflection coefficient is shown in Figure 4.2.

Next, we run a simulation of a model where this antenna is placed on a substrate. In the dialog box displayed after selecting Circuit->Dielectric Layers..., add a substrate as shown in Figure 4.3.

Figure 4.4 shows a plot of the reflection coefficient of this antenna. After selecting File->Add File(s)..., add a plot of an antenna in free space (small rectangle markers) to compare, and we can see the influence of the 1.6-mm thick dielectric. We know that the resonant frequency is now near 514 MHz. When building a planar antenna on the surface of a dielectric, the resonant frequency shifts to a lower frequency.

4.1.2 Wavelength Shortening Effect of Dielectrics

When building an electric field detection type dipole antenna on the surface of a dielectric substrate, the velocity of the traveling electromagnetic waves slows

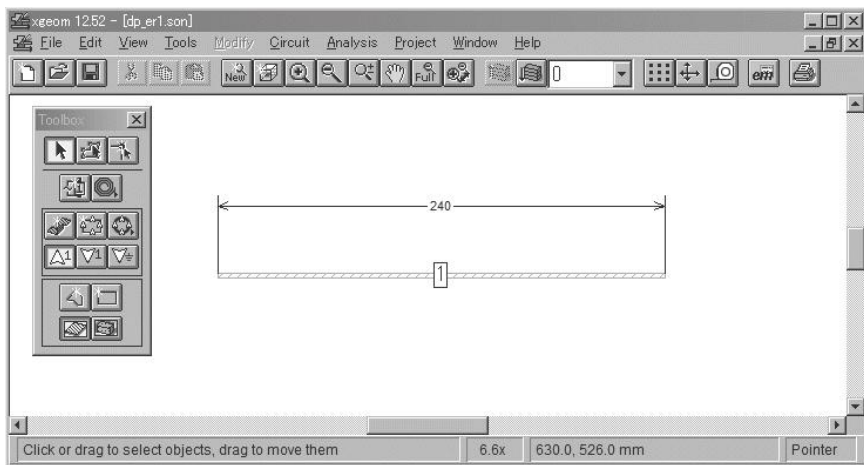


Figure 4.1 The dipole antenna model from Chapter 1 has been modified for this example. File: dp_er1.son.

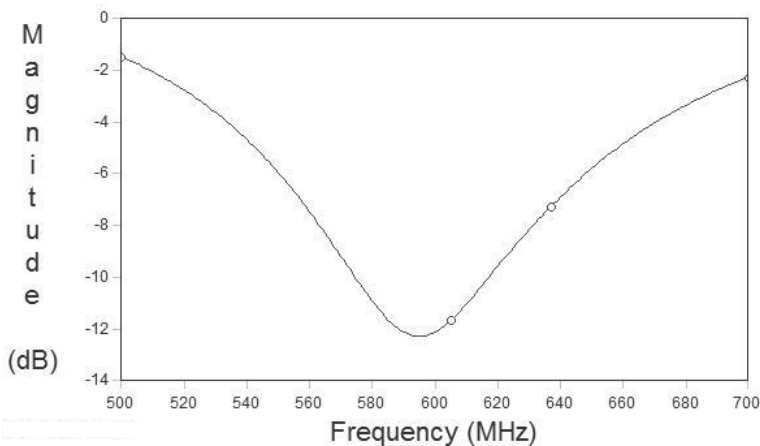


Figure 4.2 The reflection coefficient magnitude of the dipole antenna suggests that it resonates at around 600 MHz.

down as compared to free space. It could be thought of as being due to the extra time it takes for the electric field of the wave to push and pull on the electrons in the material of the dielectric (displacement current). Magnetic materials similarly lower the resonant frequency.

Figure 4.5 shows an antenna in free space and one embedded in a dielectric. If the traveling waves race from the feed point, the wave on the left wins

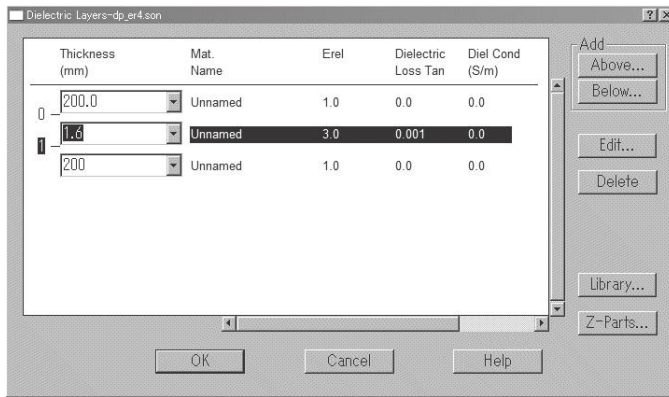


Figure 4.3 Add a substrate in the Dielectric Layers dialog box. File: dp_er3.son.

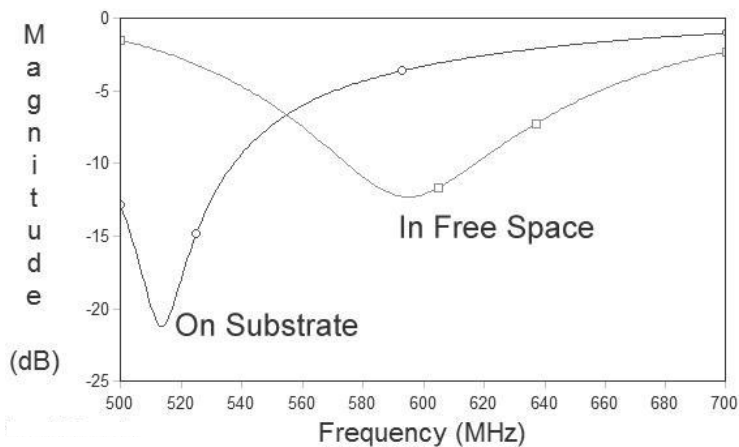


Figure 4.4 A plot of the reflection coefficient of the dipole antenna on a substrate suggests that the resonant frequency shifts down.

because it does not have to force its way through the giant cloud of electrons that are held in place by the atoms of the dielectric on the right.

In order that both antennas resonate at the same frequency, the traveling waves on each antenna must reflect back from the ends of the dipole at the same time. Thus the antenna on the right must be shortened as indicated by the dashed lines.

As just described, it is possible to miniaturize antennas by using this wavelength shortening effect. When most of the energy of the antenna's electric field is within a dielectric, the coefficient of wavelength shortening is represented as

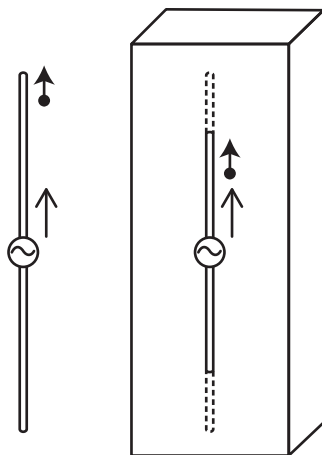


Figure 4.5 A dielectric substrate shortens wavelength, lowering the resonant frequency.

$1/\sqrt{\epsilon_r}$, where ϵ_r is the relative permittivity of the dielectric. More typically, we build the antenna on one side of a thin substrate. In this case, some of the electric field energy must push through the substrate dielectric, and some is in free space. The wavelength shortening effect is not as strong. The situation is similar for a microstrip line, the same wavelength shortening/velocity slowing occurs. The coefficient of wavelength shortening is represented as $1/\sqrt{\epsilon_{eff}}$, where ϵ_{eff} is the effective relative permittivity of the dielectric.

For a given length of microstrip line, the effective relative permittivity is $\epsilon_{eff} = C/C_0$, where C is the capacitance between the line and ground with dielectric in place, and C_0 is the capacitance with air as dielectric everywhere.

4.1.3 Investigating Wavelength Shortening Effect in an MSL

In this section, we simulate a straight microstrip line (MSL). A microstrip line is a printed circuit line on a substrate. The bottom side of the substrate is the ground (return current flows on the ground; see Figure 1.2.) There is no substrate above, only air. For our MSL, the size of the substrate is 30×30 mm, and the wire width is 1 mm, the thickness of dielectric is $300 \mu\text{m}$ and the permittivity is 4.8. The characteristic impedance of this line is not 50Ω

The MSL has been studied in detail over the years. Accurate and detailed equations have been developed to calculate its characteristic impedance. Figure 4.6 shows the AppCADTM (available at <http://www.hp.woodshot.com/>) application by Agilent Technologies, Inc., which calculates the characteristic impedance using these equations. We input the line width, the dielectric substrate thickness, and the relative permittivity.

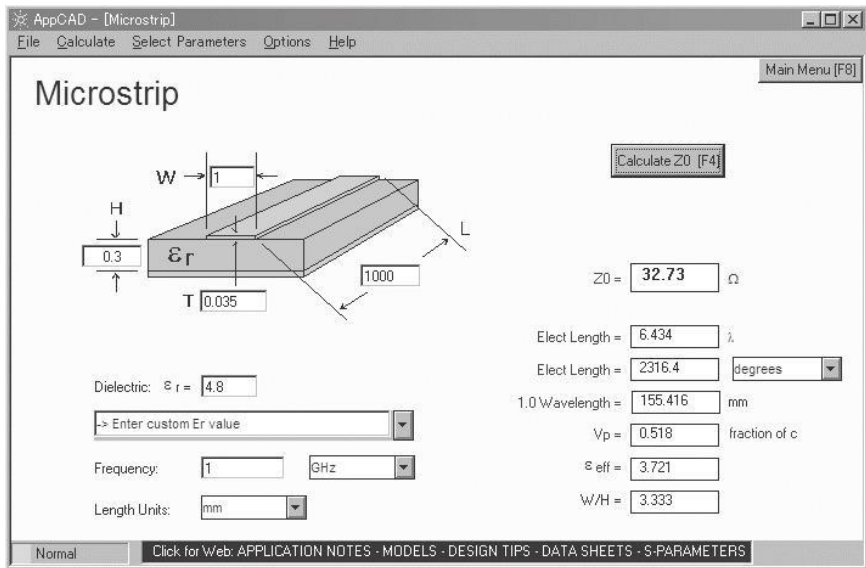


Figure 4.6 AppCAD calculates the characteristic impedance of an MSL based on closed form approximations that are calculated quickly.

Using a model of an MSL in Sonnet Lite (Figure 4.7), we see the characteristic impedance at 1 GHz is 34.5Ω as shown in Figure 4.8.

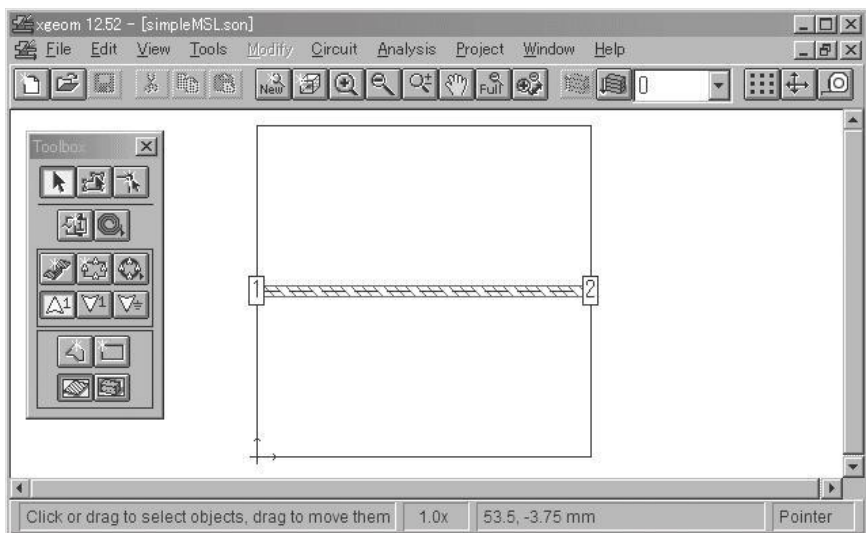


Figure 4.7 Sonnet Lite model of an MSL. File: simpleMSL.son.

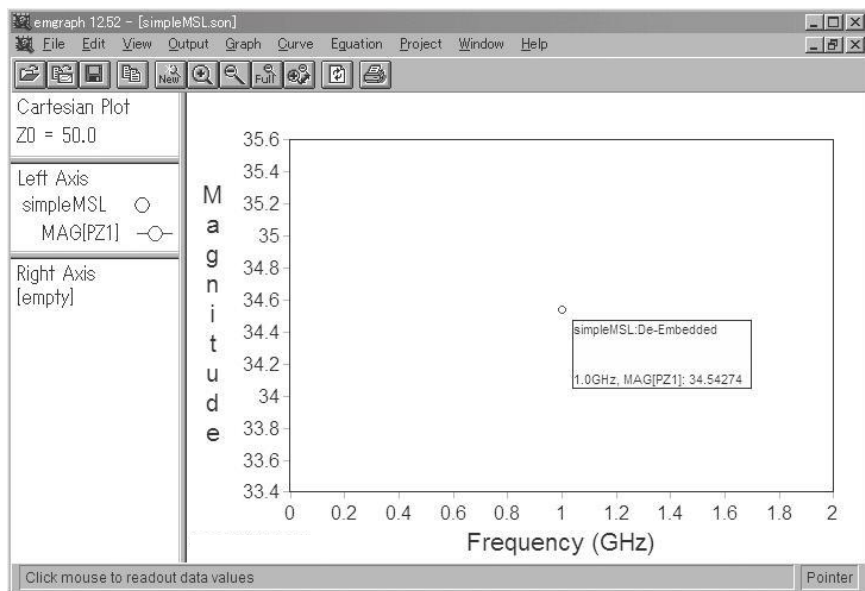


Figure 4.8 Characteristic impedance at 1 GHz is 34.5Ω .

The broadband simulation result is shown in Figure 4.9. Most electromagnetic field simulators calculate S -parameters assuming that the transmission lines connected to all ports have a 50Ω characteristic impedance. Because this MSL has a characteristic impedance of 35Ω , it looks like its reflection coefficient (S_{11}) is large over the whole range due to the impedance mismatch. However there are some specific frequencies with no reflection and it seems to be well matched!

As is well known, at these zero reflection frequencies the line lengths are an integral multiple of a half-wavelength. The physical length of the line is 30 mm. We see that the first zero reflection frequency is 2.55 GHz. In free space, one wavelength at 2.55 GHz is 3×10^8 m/s divided by 2.55×10^9 Hz, which is just under 120 mm. One half-wavelength is 60 mm. But the line is only 30 mm long! Thus, the wavelength on the MSL is shorter by half as compared to free space. This means that ϵ_{eff} for this MSL is about 4 (1 over the square root of 4 is one-half).

4.2 Fundamentals of an Inverted L Antenna

Let's start working with a more complicated antenna. If we don't have room to put a nice straight vertical monopole, let's just bend it over.

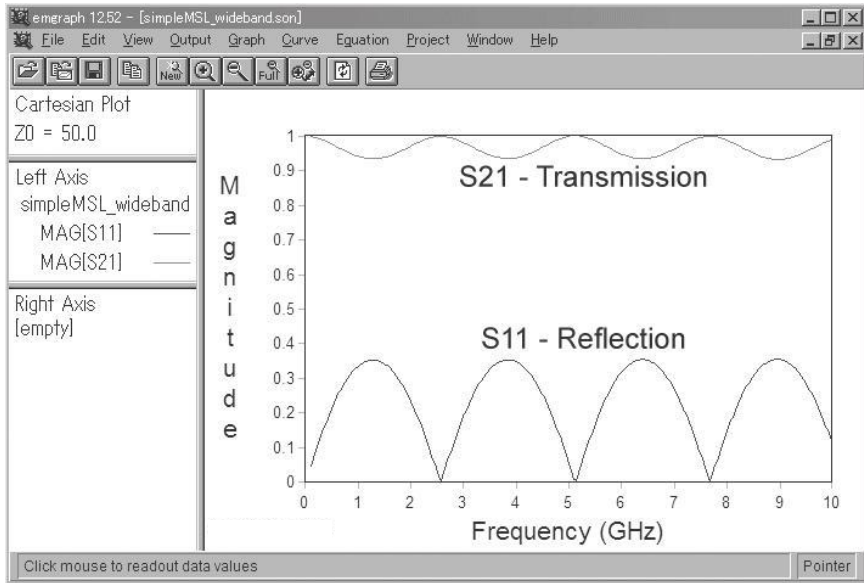


Figure 4.9 S-parameters for the MSL from 0.1 to 10 GHz. File: simpleMSL_wideband.son.

4.2.1 What Is an Inverted L Antenna?

An inverted L antenna is like a monopole antenna that has been bent near the ground so that it looks something like the letter L on its side. As shown in Figure 4.10, the directions of the currents on the antenna and on the ground are opposite each other. The more that the horizontal element approaches the ground, the more strongly it couples to the ground. As it gets very close to the ground, it is just like the MSL (see the above section), except that for the MSL, the ground current flows in a good conductor. Usually the ground for the inverted L is earth, and that is usually a resistive ground. It can be viewed in terms

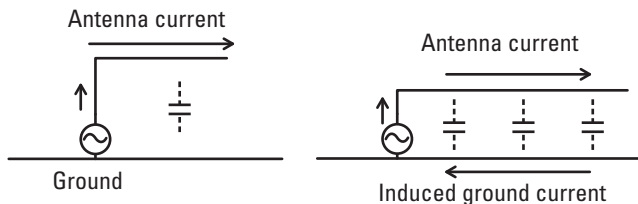


Figure 4.10 The inverted L antenna. Electromagnetic coupling between the antenna and a ground is strong.

of an equivalent capacitor to ground that is distributed along the length of the horizontal part of the antenna.

So, when the antenna is close to the ground, this structure closely resembles the MSL and this makes it difficult to radiate the electromagnetic waves. The MSL is designed to avoid unwanted radiation by being close to the ground return current. Remember the aluminum foil in Figure 1.2. If the distance between the line and the ground return path is small, then we have a transmission line that does not radiate. If the distance is large, then we have a loop antenna, which radiates well. So the question is, does the inverted L antenna radiate well or not?

To answer that question, we simulate an inverted L antenna and compare it to a monopole antenna. Our model of a monopole antenna (Figure 4.11) uses one side of the dipole antenna in Figure 4.1. Then we use the Sonnet side walls as a ground, just like Marconi's antenna in Chapter 2.

As we compare the modified antennas, we build models in free space with no dielectric substrates. The theoretical input impedance for a dipole in free space is 73Ω , pure real. Thus, a monopole above perfect ground should be just over 36Ω . Sonnet Lite gives us 38Ω , which is more than close enough (Figure 4.12).

Next we build a model with half of the element bent by modifying the monopole in Figure 4.11. Select Edit->Divide Polygons, and your cursor becomes a small razor knife. We left-click and drag the mouse across the center of the length of the monopole. Release the mouse button after you have crossed over the monopole. The monopole is still the same shape and size; it is just split into two polygons.

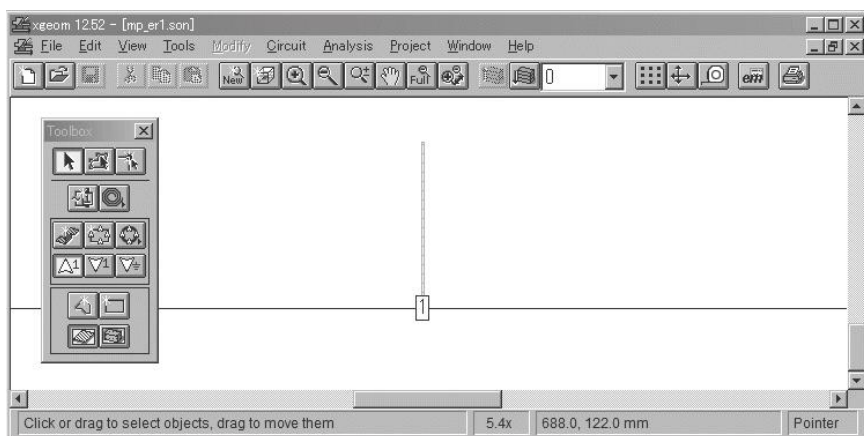


Figure 4.11 Model of the monopole antenna. File: mp_er1.son.

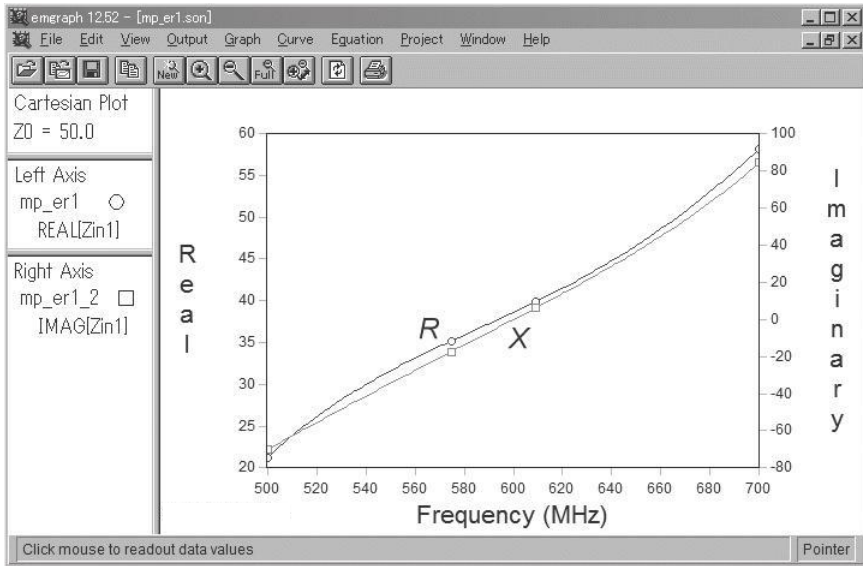


Figure 4.12 Input impedance of the monopole antenna. R is 38Ω at 600 MHz.

Next select the upper polygon by clicking inside of it. Then select Modify->Rotate... to rotate it by 90° (either direction is OK). Sometimes, the rotated polygon will not be precisely on a cell boundary. If this is the case, select the polygon again. Then select Modify->Snap To... and snap the polygon to the cell boundaries. Finally, move the rotated polygon into place. Figure 4.13 shows the

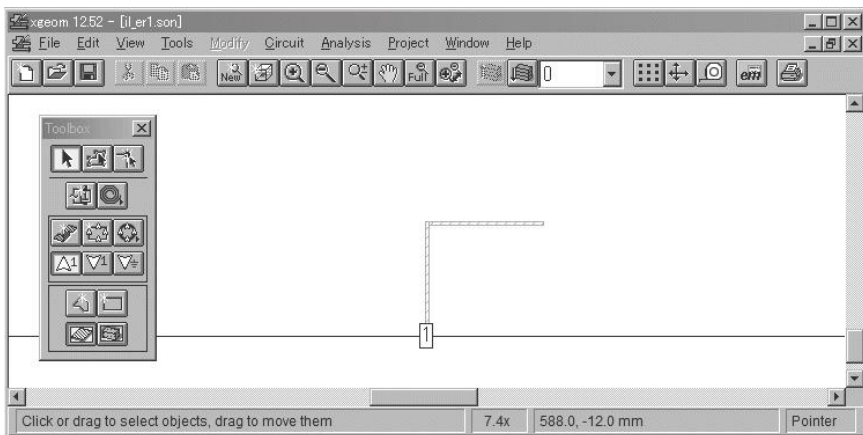


Figure 4.13 The inverted L antenna. File: il_er1.son.

inverted L antenna that we will analyze. Then select File->Save As and save it under another name.

The reactance of the monopole antenna (Figure 4.14) is 0Ω at resonance, at around 600 MHz. However the inverted L reactance is -20Ω at that frequency, which means it is capacitive.

An inverted L antenna that is designed by lengthening and lowering the horizontal element (Figure 4.15) is also based on the monopole and its total length is the same. However, because it is closer to the ground, the impedance is changed.

By changing the plotted quantity to Z_{in} (Figure 4.16) and reading the impedance, R is 7Ω , which is much too small to feed using a 50Ω coaxial cable. Note that the inverted L antenna is unbalanced, so if the impedance were 50Ω , it would be possible to connect coax directly to it. That would certainly be convenient!

Next we design a circuit that converts the impedance of this antenna ($7 - j20\Omega$) to 50Ω . The classical approach to impedance matching uses a Smith chart. However, here we use a feature included in a circuit simulator.

Figure 4.17 is a screen of S-NAP Design (MEL Inc.; this software is available only in Japanese), where we assign 50Ω to the input impedance and $7 - j20\Omega$ to the load impedance. Then it creates a matching circuit.

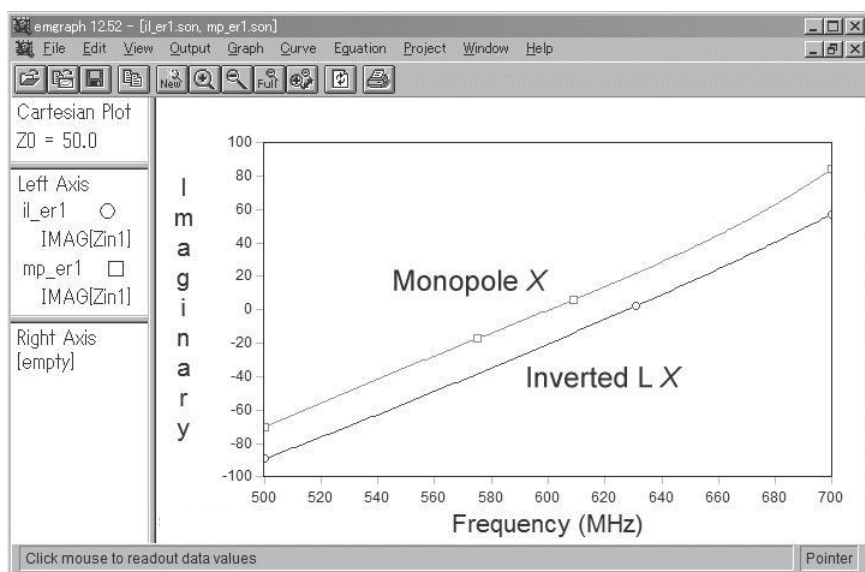


Figure 4.14 Reactance of the L antenna (small circle marker) and the monopole (small square). File: il_er1.son.

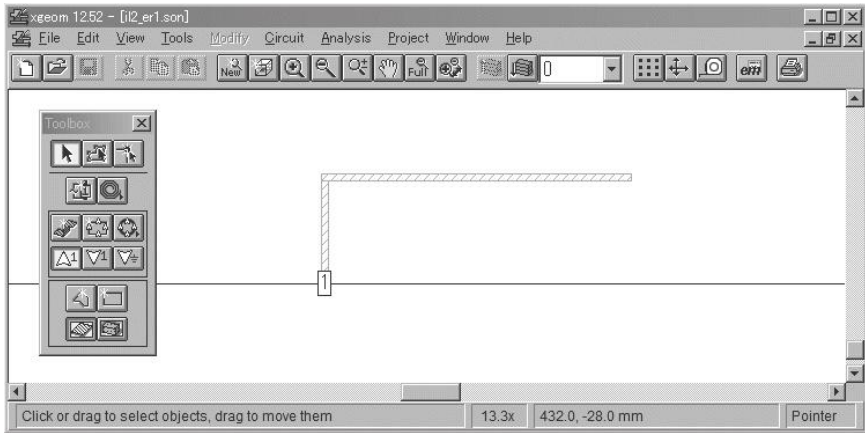


Figure 4.15 Inverted L antenna bent closer to the ground. File: il2_er1.son.

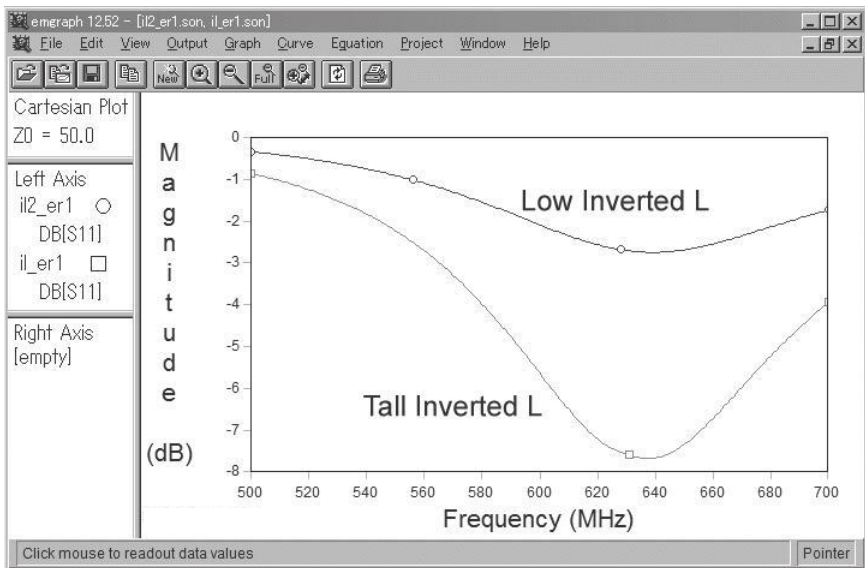


Figure 4.16 Reflection coefficient of the inverted L antenna (small circle marker).

From the S-NAP results, connecting a 10-nH coil in series between the coax center conductor and the antenna input, and then also connecting a 13-pF capacitor from the coax center conductor to the coax ground, the input impedance of the antenna should appear to be 50Ω from the coax's point of view.

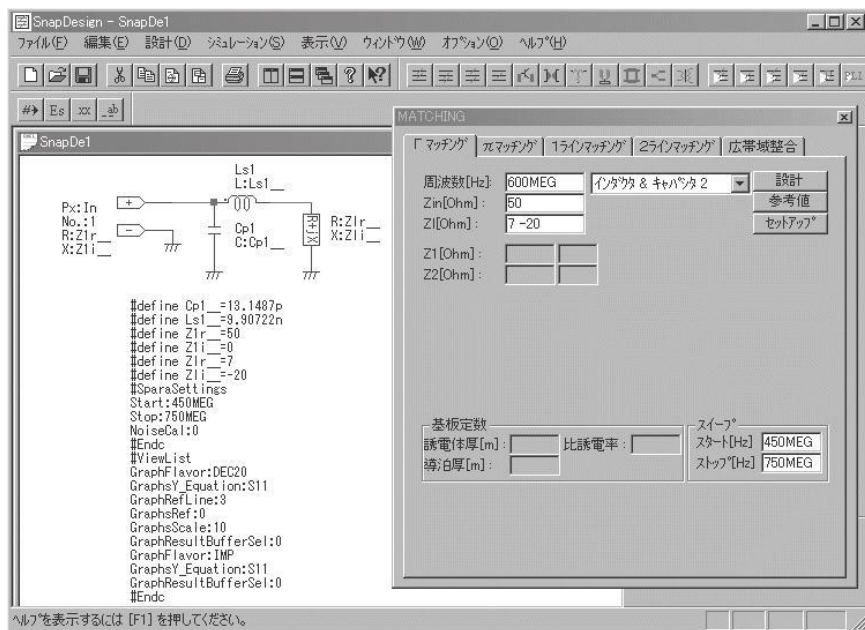


Figure 4.17 Using S-NAP Design to create a matching circuit.

To verify this, we use Netlist Project in Sonnet Lite that we previously used in Chapter 2.

Figure 4.18 is the circuit as designed. In the result of S_{11} (Figure 4.19), we see a good, if narrowband, match to 600 MHz as designed. Note that when matching widely different impedances, that bandwidth of a lossless (no resistance) matching network tends to be very small. If there is loss, the network

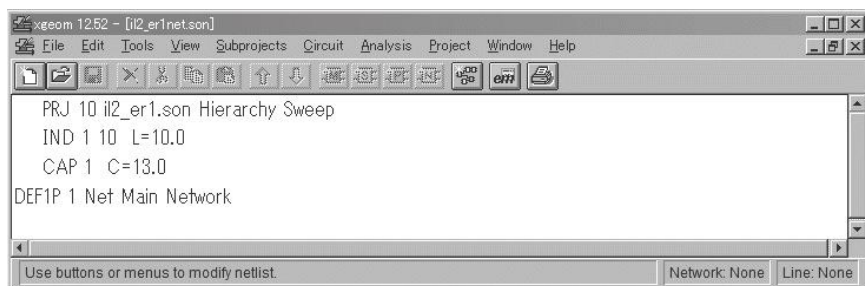


Figure 4.18 Sonnet Netlist Project verifying the matching circuit.

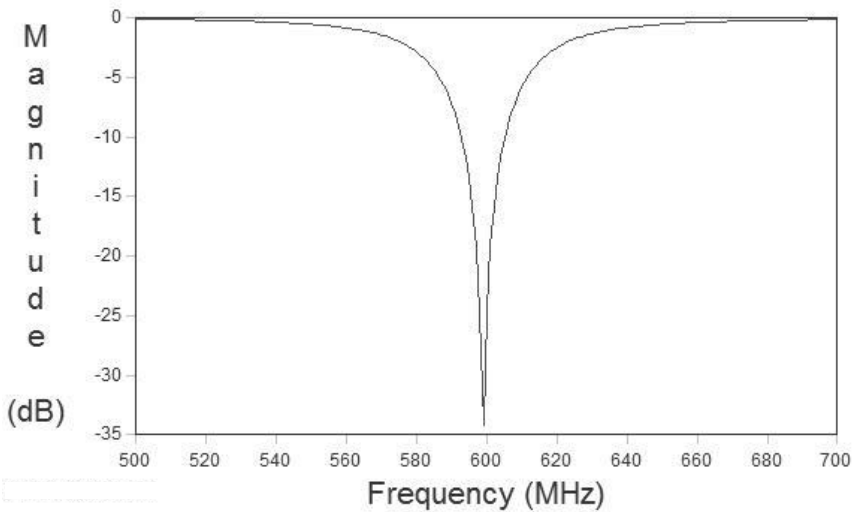


Figure 4.19 Result of S_{11} for the antenna plus matching network. It is matched at 600 MHz.

becomes more inefficient as the change in impedance levels becomes large. This is all related to something you might hear referred to as *Fano's limit*.

In an actual antenna matching network, stray inductances and capacitances (as in short wires used to connect the components together) makes it difficult to realize exact values. In addition, fabrication tolerances for the antenna mean that the calculated input impedance is different in the actual antenna. A side effect of being narrowband is that the component values are critical. The net result is that matching networks are usually built using variable capacitors and inductors (with as low loss as possible). The components are then tuned (manually or automatically) to match the actual antenna for the selected operating frequency.

Viewing the current distribution on the element (Figure 4.20), we can imagine that the area around feed point, where strong current flows, and the electrons (that form the current) are rapidly accelerated back and forth at radio frequencies, is what causes the antenna to radiate.

4.3 Fundamentals of a Patch Antenna

Perhaps the most common planar antenna is the patch antenna. You will note similarities to the common dipole, and you can apply many of the same principles you have already learned.

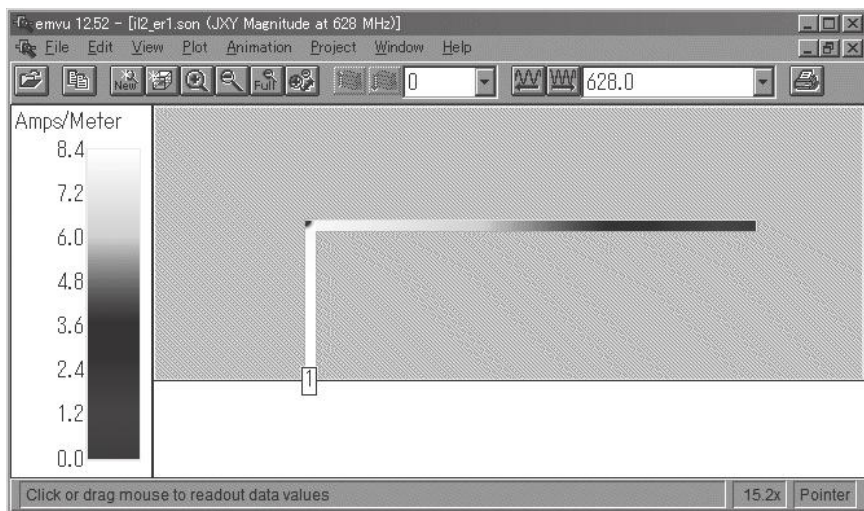


Figure 4.20 Current distribution on the inverted L antenna.

4.3.1 A Patch Antenna for Global Positioning Systems

A Global Positioning System (GPS), often used for car navigation, typically does not use dipole antennas. This is because the antenna is usually placed on the metallic surface of a vehicle. A patch antenna works here because the substrate on which the patch is fabricated separates it from the car body. So this antenna is widely used not only for car navigation systems but also for compact antennas inside electronic products.

Figure 4.21 is a simulation model with a rectangular metallic patch over a wide metallic surface (the ground plane), and the electromagnetic waves from the GPS satellites enter into the gap between the patch and its ground plane. The electromagnetic energy flows between the plate and ground like a capacitor. The receiver is excited by the voltage induced between the plate and the ground plane.

The size of the patch is designed so that one side is a half-wavelength. When there is a substrate with a high dielectric constant between the patch and its ground plane, the wavelength (and thus the antenna) is shortened, as described in Section 4.1.

Figure 4.22 is a small patch antenna for GPS by Yokowo Co., Ltd. The outside dimensions are $25 \times 25 \times 4$ mm, $20 \times 20 \times 4$ mm, and $18 \times 18 \times 4$ mm from the left.

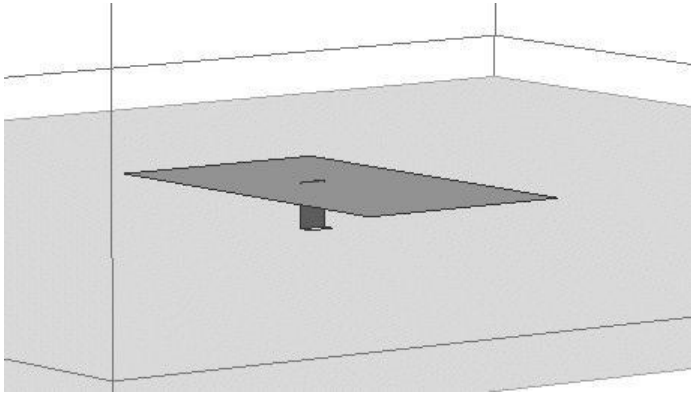


Figure 4.21 Sonnet model of a patch antenna.

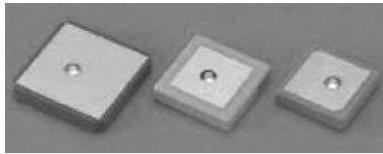


Figure 4.22 Small-size patch antennas for GPS by Yokowo Co., Ltd.

4.3.2 Electromagnetic Field Around a Patch Antenna

A patch antenna is also called a microstrip antenna (MSA). When building it on a substrate, it can be fed through an MSL. The patch can also be viewed as a microstrip line where the width of the MSL is enlarged at the end.

Figure 4.23 is a model of a patch antenna. For simplicity, we model this patch antenna with an air substrate, with a relative dielectric constant of 1.0. Then we look at a patch antenna with a real substrate. The thickness of the substrate is 1 mm, the cell size is 1×1 mm and the substrate is set to 200×200 mm (x and y Box Size). The dotted line can be set in the Box by checking Symmetry (Circuit->Box). When this option is selected, only the top half of the circuit requires analysis time. A mirror image of the top half is automatically included in the analysis at no extra cost. This can make an analysis go considerably faster and it cuts the required memory to up to one quarter.

The current on the patch antenna at 4 GHz, where S_{11} is minimized, is shown in Figure 4.24. It shows that electromagnetic waves travel along the MSL and a symmetrical pattern of surface current is generated on the symmetrical patch.

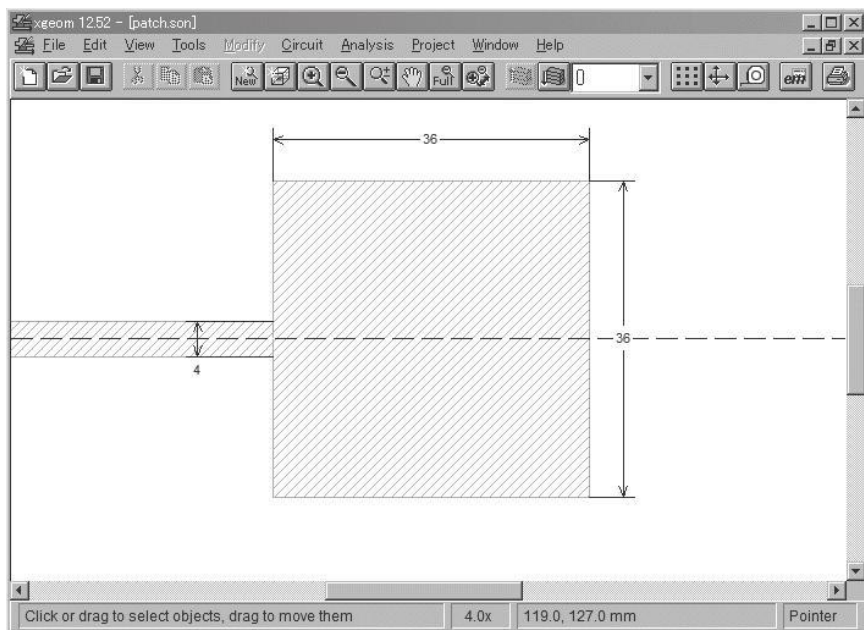


Figure 4.23 Model of a patch antenna. The patch is 36×36 mm.

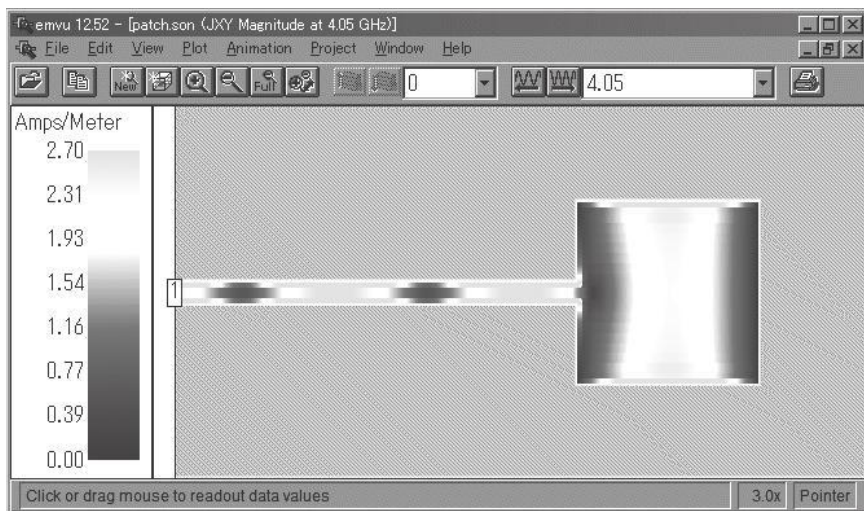


Figure 4.24 Surface current on the patch antenna at 4 GHz. File: patch.son

At high frequencies, the current on a line flows most strongly on the edges. This is called the edge singularity. You can also view the sinusoidal standing

waves on the line. The patch itself is a standing wave antenna, just like the dipole (Figure 3.2). You can see the standing half-wavelength sinusoidal distribution on the patch.

Normally, we would see no standing waves on the feed line. This is indeed the case when the antenna is well-matched to the feed line. In this case, however, because we see strong standing waves on the feed line, we know that this antenna is not well-matched to the feed line. The input impedance of the antenna is significantly different from the characteristic impedance of the feed line.

Returning to the patch, remember the half-wavelength dipole antenna; we know that the current on the patch in Figure 4.24 seen from the side forms a sine wave, as shown in Figure 4.25. And when it resonates like this and a standing wave is generated, the voltage distribution is shown with the dotted line, and we have high voltage/low current at the edges of the patch.

We can imagine certain aspects of electromagnetic field around the patch from this diagram. Figure 4.26 shows the electric field distribution between the patch and the ground plane.

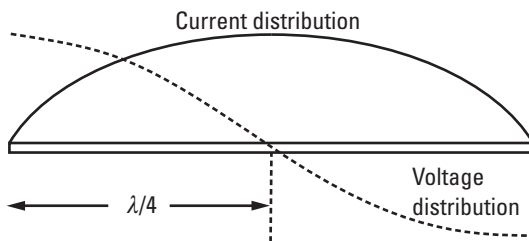


Figure 4.25 Current and voltage distribution on a patch.

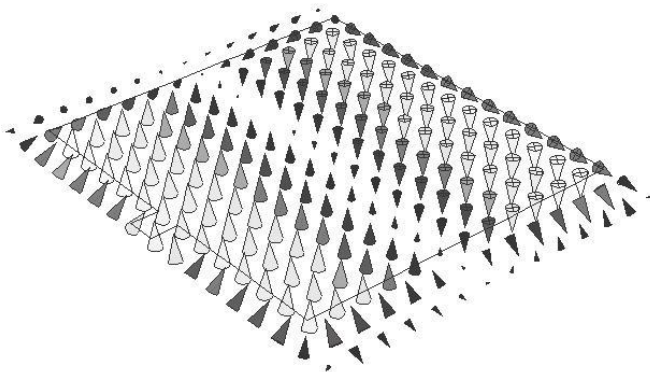


Figure 4.26 Electric field distribution between the patch and the ground plane.

We see that the electric field vectors distribute vertically on the patch and the ground surface. Each small cone represents the electric field vector in 3-D. Here the direction of the vectors are reversed on opposite sides of the patch. The voltages at the edges of left side and right side are opposite, just as shown in Figure 4.25. Patch antenna radiation can be understood as a result of the strong electric fields, which bulge out from both high voltage edges into free space.

In textbooks on antennas, this bulging electric field is modeled as a discontinuity in the electric field which can, in turn, be modeled as a flow of fictitious magnetic current. Magnetic current is the flow of magnetic monopoles, the magnetic counterpart of electrons. Magnetic monopoles and magnetic current have never been observed in nature, but replacing the antenna with a magnetic current in the mathematics used in the textbooks gives first-rate results.

4.3.3 Determining Dimensions of a Patch Antenna

Figure 4.27 is a graph of the reflection coefficient of the patch antenna in Figure 4.22. It resonates at around 4 GHz; however it is about -4 dB, so the reflection, and thus the impedance mismatch, is a bit large.

For this illustrative example, we set the relative dielectric constant of the substrate to 1.0 and we are going to try to make a patch antenna that resonates at 4 GHz. At that frequency, a half-wavelength is $3 \times 10^8 / (4 \times 10^9 \times 2) = 37.5$ mm. However, when we use that dimension, the patch resonates at around

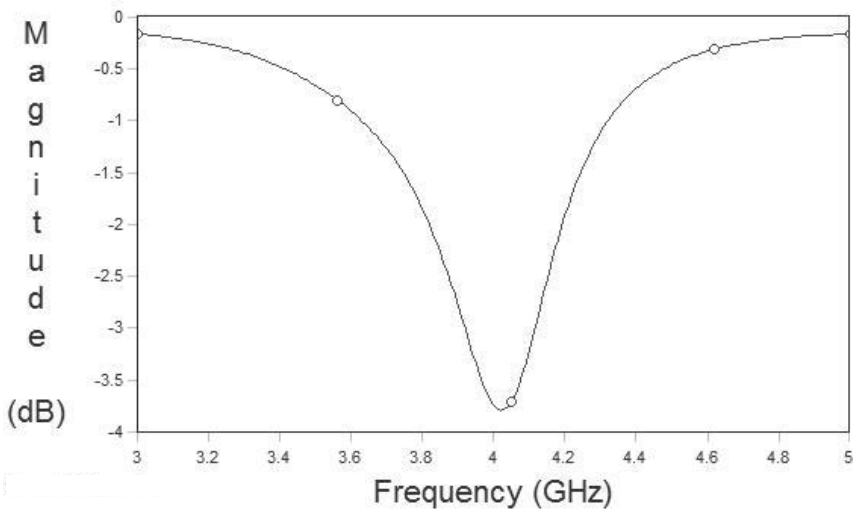


Figure 4.27 Reflection coefficient of the 36-mm square patch antenna in Figure 4.23, including the MSL feed line.

3.8 GHz. After trying several adjustments of the patch size, we obtained the reflection coefficient shown in Figure 4.27, with the patch 36 mm on a side.

Note that Figure 4.27 shows S_{11} normalized to 50Ω . Even if the characteristic impedance of the feed line is 50Ω , the impedance seen from the feed point of the patch is significantly different from 50Ω . The characteristic impedance of the fairly long MSL feed is actually just under 60Ω , making the situation very complicated indeed. It would be nice if we could remove the complexities of the feed line.

Sonnet can exactly remove the effect of the MSL feed line and determine the impedance seen right at the feed point of a patch. First, get the dialog box shown in Figure 4.28 by selecting Circuit->Ref. Planes/Cal. Length..., and then selecting Fixed and clicking on the mouse icon. The cursor changes to +. Then click exactly on the patch feed point and click OK. Figure 4.29 shows that the reference plane is assigned. Sonnet now calculates and plots S -parameters as seen from the location of the sharp end of the arrow. The effect of the MSL feed line is perfectly removed.

Figure 4.30 shows the impedance of the patch antenna seen from the reference plane (the end of the arrow). We see the peak value of R is 288Ω at 3.93 GHz. In addition, as we see in the plot, X is 0Ω at the same frequency, indicating resonance. Thus, with the effect of the feed line removed, the resonance has moved down slightly. The 288Ω input impedance shows why we see a poor R in Figure 4.27. The input impedance is too far from 50Ω . If we were to use an MSL with a characteristic impedance of 288Ω , and design our transmitter and receiver to properly drive that impedance, we would have no problem. Well, at least until we actually tried to design such a high impedance MSL. The width

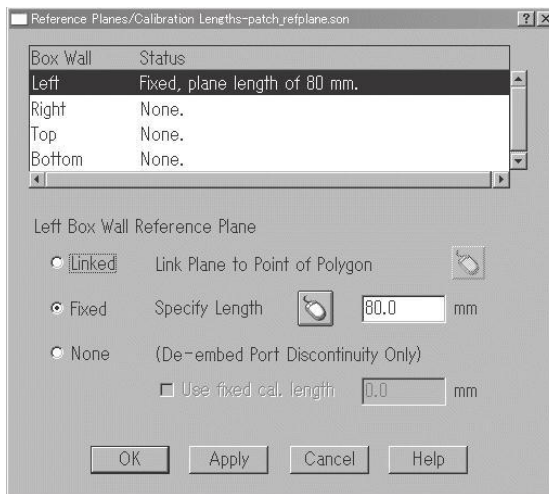


Figure 4.28 Dialog box to set the reference plane.

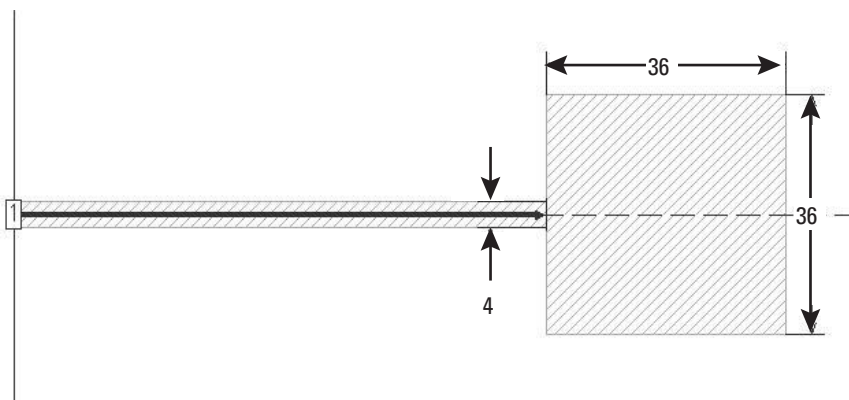


Figure 4.29 The reference plane is located at the sharp end of an arrow. File: patch_ref-plane.son.

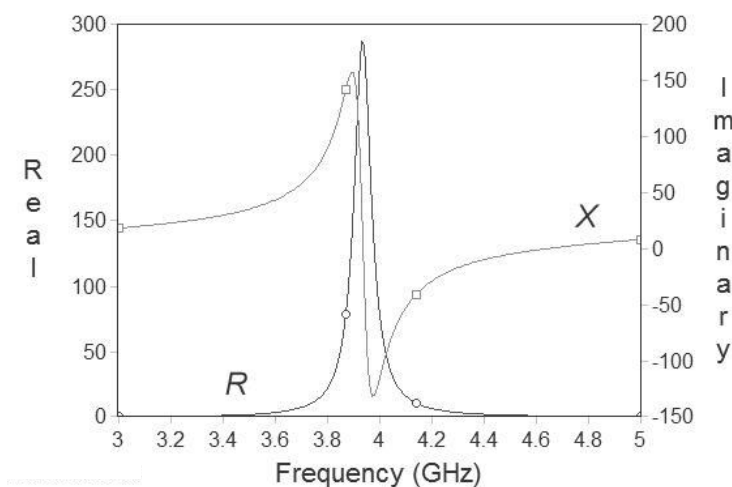


Figure 4.30 Impedance of the 36-mm square patch antenna seen from the reference plane. The real part uses circle markers, while the square markers indicate the imaginary part.

of the line would be so tiny, it would be very resistive and lossy, if it could even be fabricated.

Our initial, very simple calculation of the required dimension for a 4-GHz resonant frequency of 37.5 mm did not include the electromagnetic “fringing” fields bulging from the edges of the patch. The resulting resonant frequency was too low. So we shrunk the antenna and found resonance at about 4 GHz with a size of 36 mm. Then, we removed the effect of the MSL feed line and we

found that the antenna was really resonating at 3.93 GHz. We need to shrink the patch even more, but just a little bit. For this example, we change the cell size to 0.5 mm and shrink the patch only in the x direction to 35.5 mm.

Figure 4.31 shows the impedance of the patch antenna seen from this reference plane. The peak value of R is 285Ω , and X is almost zero at 3.98 GHz. Close enough.

4.3.4 A Patch Antenna on a Substrate

A patch antenna is typically built on a substrate with dielectric materials. So, to make our example more realistic, we change the bottom dielectric layer from air to FR-4 in Sonnet Lite.

Select Circuit->Dielectric Layers... and in the displayed dialog box double-click the lower layer. In the new dialog box (Figure 4.32) click [Select dielectric from library...] in the upper left and select FR-4. The relative permittivity is set to 4.9 and the loss tangent is 0.025; these values are already registered in the library. (Note that in practice, the dielectric constant of FR4 is not well controlled due to manufacturing variability.)

A simple first guess for the resonant frequency of a patch antenna is

$$f = \frac{3 \times 10^8}{2d\sqrt{\epsilon_r}}$$

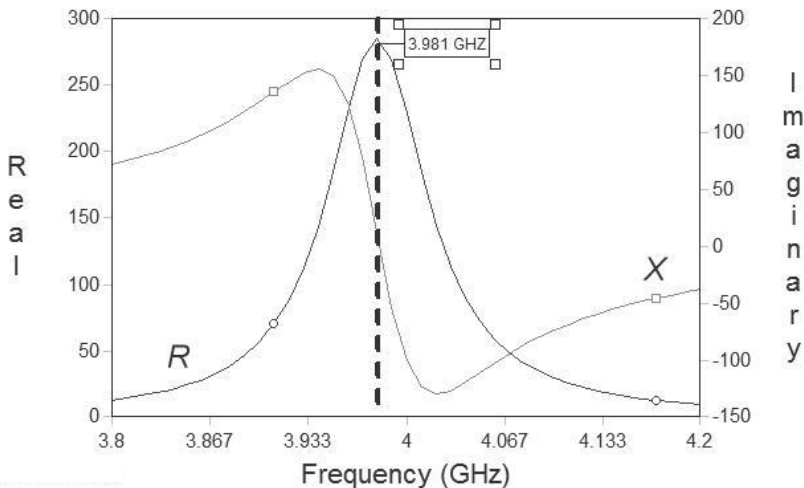


Figure 4.31 Impedance of the patch antenna with one side shortened to 35.5 mm.

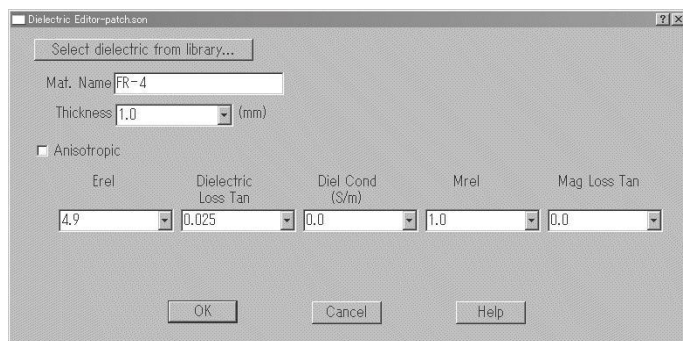


Figure 4.32 Dialog box to set values for the dielectric substrate. File: dp_er3.son.

where d is the size of the patch along the direction that the sinusoidal standing wave will form, and ϵ_r is the relative permittivity of dielectric. This equation assumes no fringing fields, and that most of the electric field is in the substrate. The error of both of these assumptions tends to cancel out. So this simple expression can actually give a pretty good answer.

When we solve the equation for d , given a resonant frequency of 4 GHz, we find that the patch should be 16.9 mm on a side.

$$d = \frac{3 \times 10^8}{2f\sqrt{\epsilon_r}} = \frac{3 \times 10^8}{2 \times 4 \times 10^9 \sqrt{4.9}} = 0.0169$$

Figure 4.33 shows the impedance when we change the x dimension of the patch in Figure 4.29 to 17 mm. The y dimension was set to 18 mm, but since the length of the current distribution half sine wave is along the x dimension, the y dimension is not critical. With the reference plane (tip of the arrow) at the feed point, R is 152Ω and X is almost zero at 3.91 GHz.

Because the resonant frequency is a little low, the patch is a little too long in the x dimension. Let's see what happens when we make the x dimension of the patch 1 mm shorter, or 16 mm. After EM analysis, the resonant frequency goes up to 4.14 GHz (patch2_FR4.son).

We see from these two analyses that a change in 1 mm of the critical patch dimension results in a resonant frequency change of 0.23 GHz ($= 4.14 - 3.91$). We want the dimension required for a resonant frequency of exactly 4 GHz. In other words, we want the resonant frequency to be 0.14 GHz less than what we get with a 16-mm patch. So the patch dimension should be $16.0 + 0.14 / 0.23 = 16.6$ mm. Of course, when we fabricate the antenna in production, there are fabrication tolerances. If we know the fabrication tolerances, we can calculate how much the resonant frequency will vary due to fabrication tolerances.

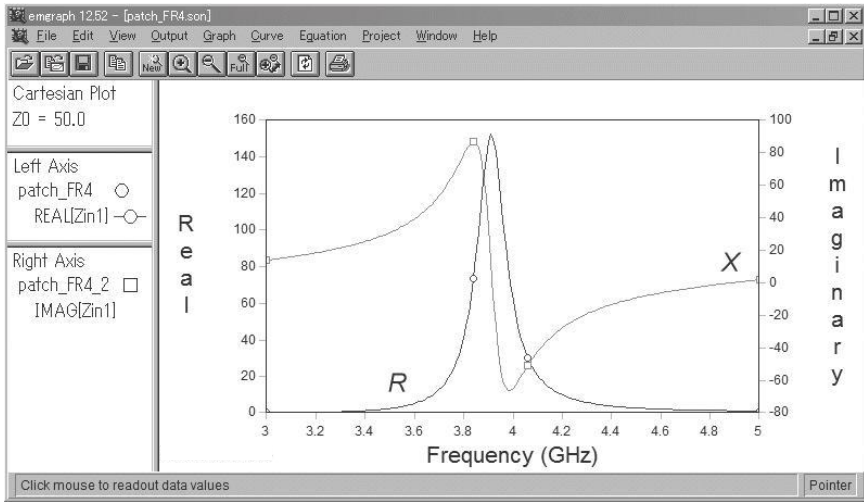


Figure 4.33 Impedance of the patch antenna on a FR-4 substrate. File: patch_FR4.son.

At this point, a common reaction is that if we want a really tiny antenna for a really low frequency, why not just use a really high dielectric constant substrate? The problem is that a high dielectric constant substrate keeps all of the electric field inside the substrate. There is nothing left over to radiate. So, the radiation efficiency is very low. Yes, indeed, we would see a nice, tiny, well-matched patch antenna, but all it does is heat up the substrate instead of radiate energy. For a patch antenna, radiation efficiency is increased by having thicker substrates with lower dielectric constants. There is no free ride.

4.3.5 Matching Method 1

Connecting the feed line right at the edge of the patch gives a really high input impedance. The impedance is so high that it cannot in practice be fed by an MSL. Fortunately, there are other ways to feed the patch with a MSL that matches nicely to 50Ω .

Notice that Figure 4.34 is similar to Figure 4.25. At the center of the patch, the voltage is zero. This means that the impedance is zero.

At the edge of a patch, we have high voltage and high input impedance. It would seem that there is a place somewhere between the edge and the center where we can connect a 50Ω MSL and get a good match, something like in Figure 4.35. To demonstrate, we return to our first example, which has an air substrate.

Now draw the patch with three separate rectangles as shown in Figure 4.35. The slot width inside the patch is 1 mm and a strong current flows along

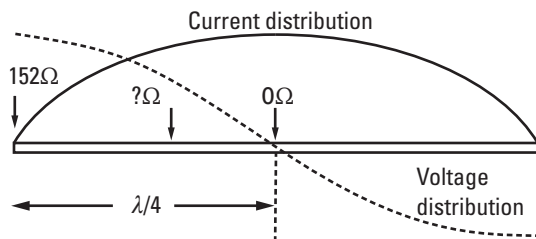


Figure 4.34 Current and voltage distribution on a patch.

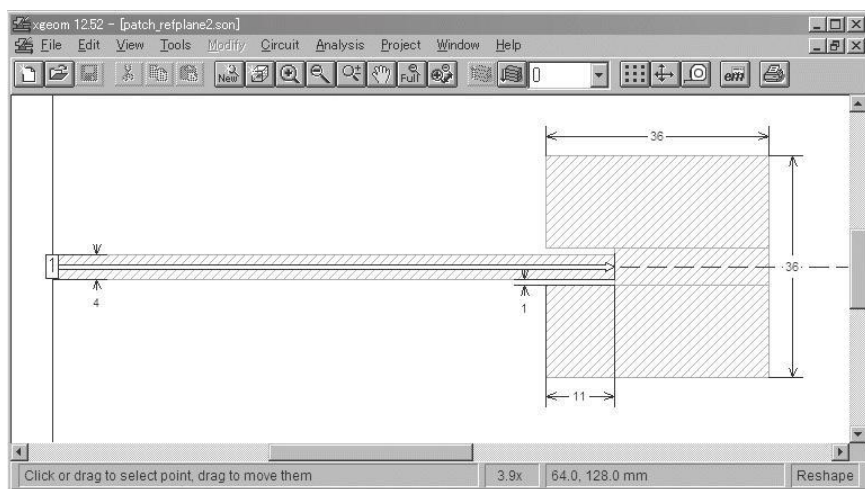


Figure 4.35 Feed position inserted into the patch. File: patch_refplane2.son

these edges. This modifies the current distribution of patch but the overall standing wave on the patch does not change significantly.

When we change the position of the feed point (by changing the 11-mm dimension shown in Figure 4.35), we will set the position of the reference plane to follow along. We will do multiple simulations, changing the depth of the feed point until we realize the desired match to 50Ω . Normally, we would want to leave the reference plane at the exact location where we will eventually connect the 50Ω feed line. In this case, for demonstration purposes, when we change the depth of the feed point inside the patch, we will also change the reference plane to the new feed point location. To do this, we change the reference plane to “Linked” (Figure 4.36). Click on the mouse icon, then click on the vertex in the patch polygon to which you want the reference plane tied. Now, when you move that point, the reference plane moves right along with it.

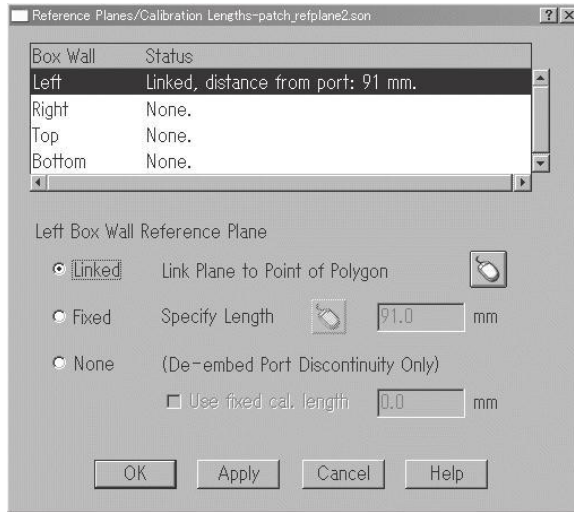


Figure 4.36 Linking the reference plane to move whenever a given vertex in the patch antenna moves.

When the depth of the feed point is 11 mm, the input impedance, R is almost 50Ω (Figure 4.37). This plot shows the impedance of feed point of a patch seen from selected the reference plane.

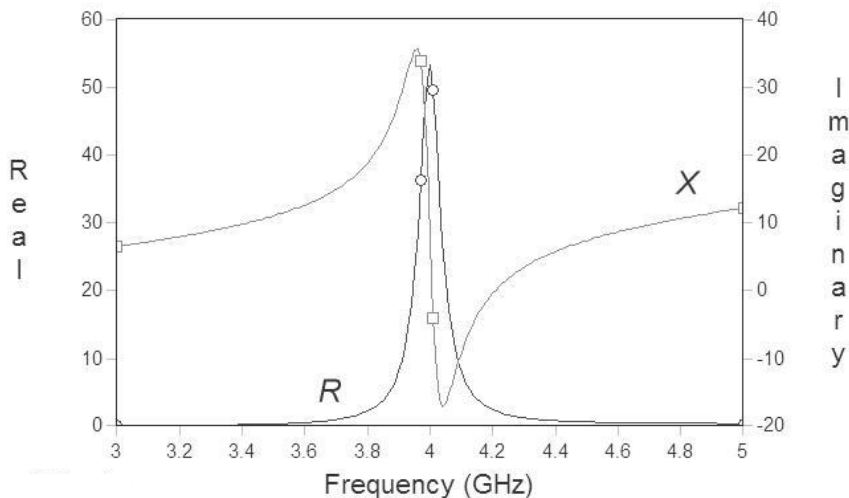


Figure 4.37 Input impedance of the patch antenna when the slot length inside the patch is 11 mm.

4.3.6 Fine Adjustment of a Feed Point

Figure 4.38 shows the patch antenna we investigated above, on FR-4, modified for a 50Ω feed. We changed the Cell Size to 0.5×0.5 mm and the Box to 80×80 mm. The slot affects the resonant frequency too, and it resonates at around 4 GHz when the longitudinal length of the patch is 17.5 mm.

Figure 4.39 is a plot of the reflection coefficient suggesting resonance at 3.95 GHz. The input impedance is $49 + j0\Omega$. The apparent resonant frequency is not quite 4 GHz. We reduce longitudinal length of the patch by 0.5 mm and the resonance shifts up by 0.12 GHz to 4.07 GHz. Knowing that changing the patch side gives 0.12 GHz of change in resonant frequency for every 0.5 mm of change in dimension, we can quickly tune the patch to the exact desired resonant frequency. Recall that a pure real input impedance is required for determination of resonance.

Figure 4.40 shows a portion of the MSL that is used as a feed line for the above patch antenna. To create this model, after saving the above patch antenna, patch_FR4_50ohm.son, in Figure 4.38 under another name (we used msl.son), we delete the patch polygons and then shorten the line.

This is an open end and all electromagnetic waves are all reflected, so the circuit itself is of little use. However, Sonnet Lite still calculates the characteristic impedance. Figure 4.41 shows the results by selecting Data Type as Port

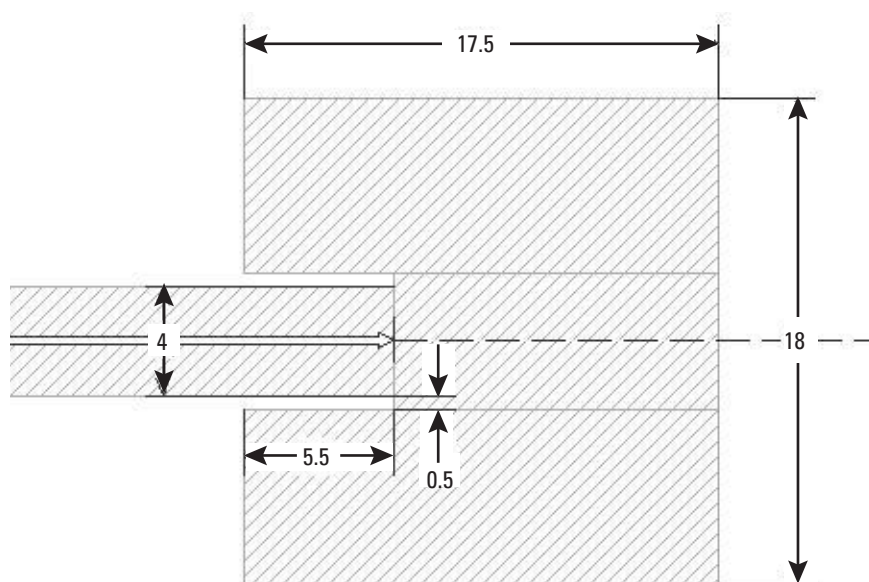


Figure 4.38 Matched feed point for the patch antenna on FR-4.

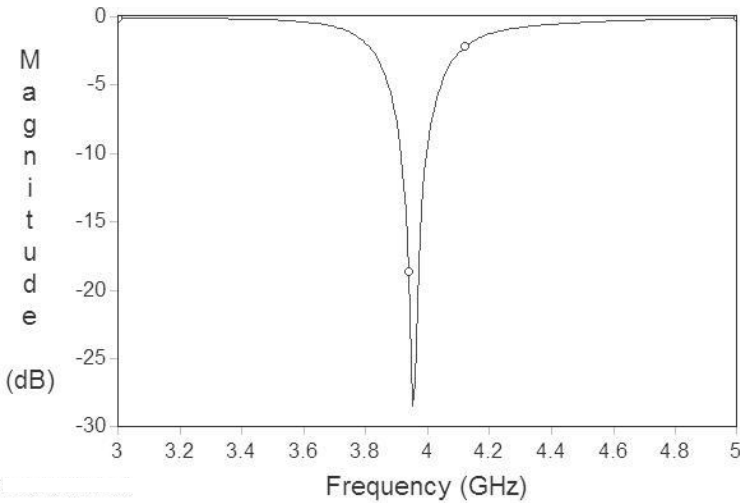


Figure 4.39 Reflection coefficient of the matched antenna on FR-4.

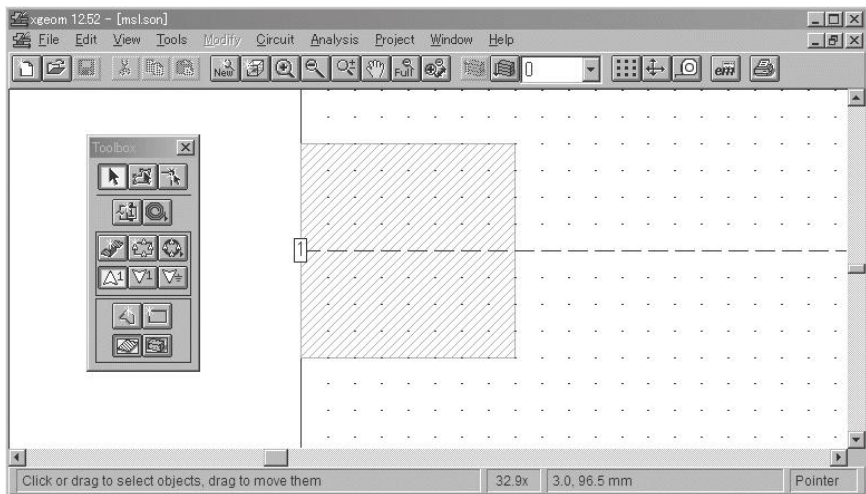


Figure 4.40 A short portion the feed line MSL on FR-4.

Z_0 for the plot. The characteristic impedance of this line is 29.6Ω , so we must adjust the line width. Narrowing the line increases the impedance.

To double-check the calculated characteristic impedance (always good engineering practice), we use AppCAD, which was shown in Figure 4.6. According to AppCAD, the characteristic impedance of the MSL in Figure 4.40 is

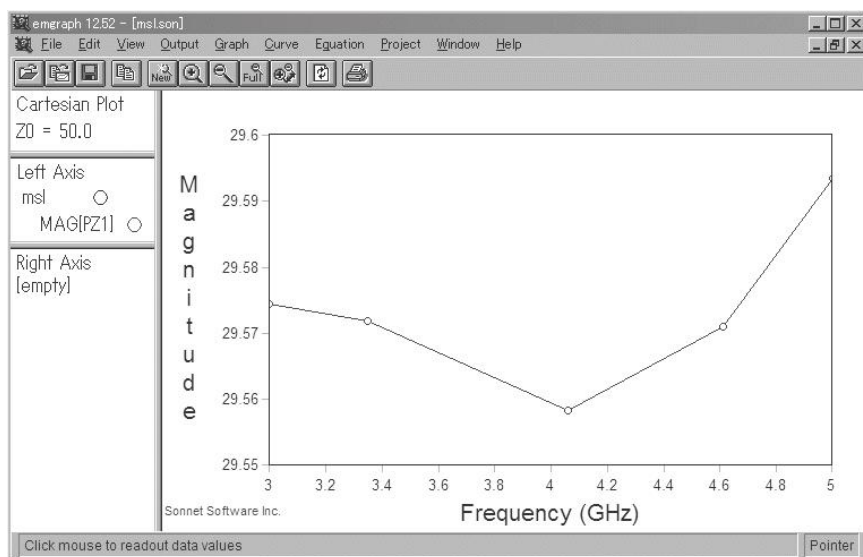


Figure 4.41 Characteristic impedance of the line as calculated by Sonnet Lite.

29.1 Ω , in good agreement with the Sonnet result. AppCAD calculates the 50 Ω line width to be 1.76 mm.

In the model shown in Figure 4.40, we change the line width to 2 mm and the characteristic impedance becomes 47.8 Ω . This is more than close enough. So next we change the line width shown in Figure 4.38 to 2 mm, then delete the reference plane (Circuit->Ref. Planes/Cal. Length..., select "None" for the Left reference plane) and find the impedance seen from the feed point of the left edge of the substrate. Figure 4.42 shows the modified layout and Figure 4.43 shows its impedance.

4.3.7 Matching Method 2

As we have seen above, moving the feed point of a patch allows us to match the antenna to 50 Ω . Another method to move the feed point is to feed the antenna from below rather than from the side. To build this kind of patch antenna, make a hole from under the bottom of ground plane and connect a coaxial (e.g., SMA) connector. The center conductor goes up through the substrate and connects directly to the patch.

Figure 4.44 shows an antenna model that uses this method. The patch size is the same as the antenna in Figure 4.42. We remove the MSL feed line, fill in the notch left from the old MSL feed line, and then place a Sonnet via port at the same position.

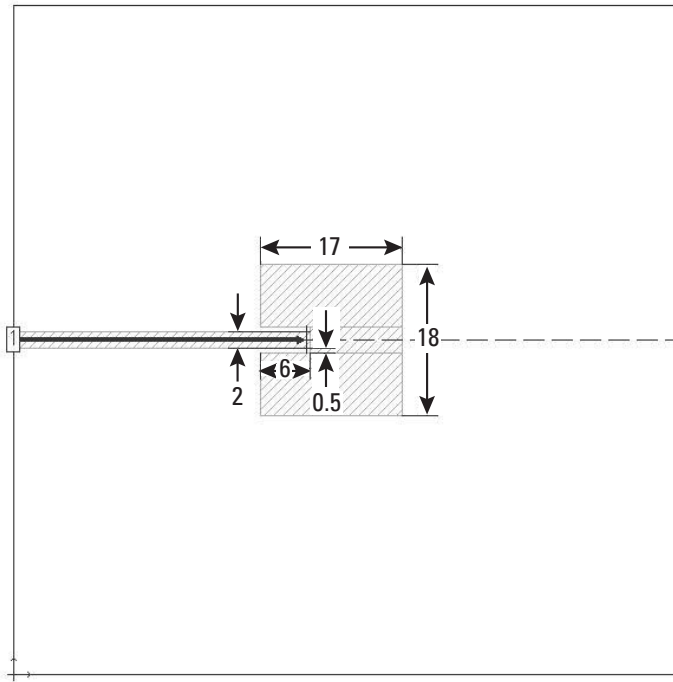


Figure 4.42 Feed line width is changed to 2 mm. File: patch_FR4_msl50ohm.son.

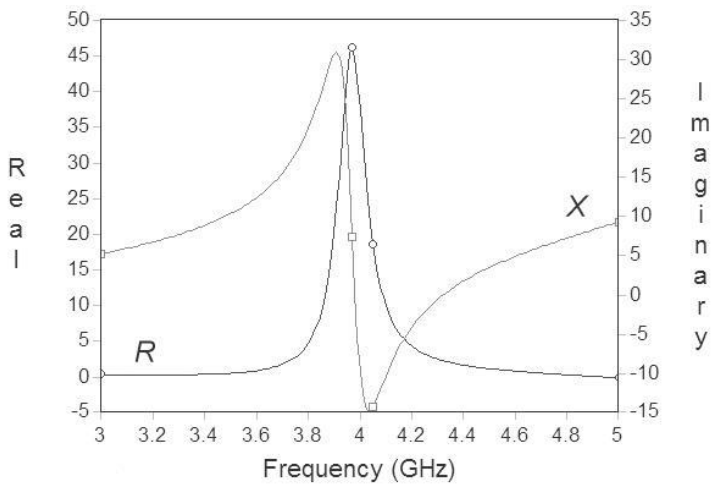


Figure 4.43 Impedance of the antenna with a line width of 2 mm. Input resistance is now matched to 50Ω .

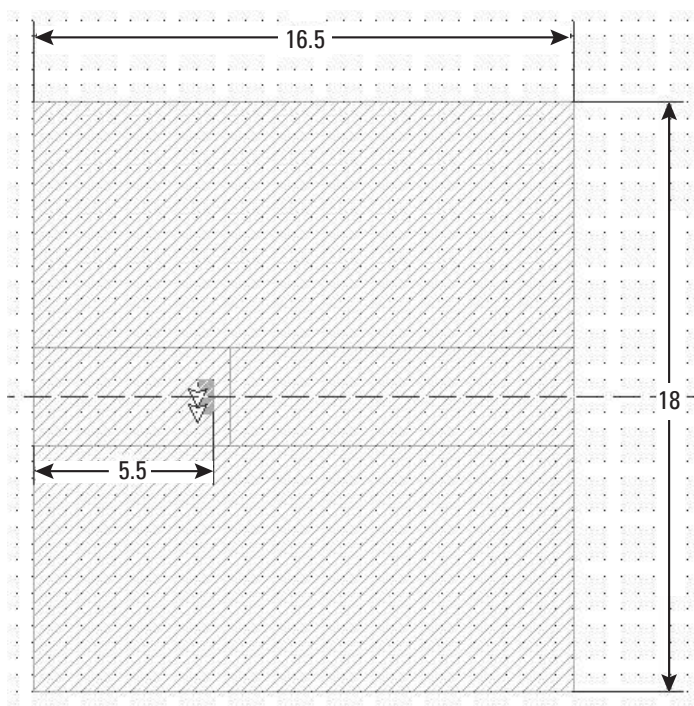


Figure 4.44 Patch antenna fed by via port, viewed on the GND level. File: patch_FR4_viaport.son.

To place a via port, first press the Ctrl key and D key at the same time, or press the down arrow key, and go down to the GND level. Next, as shown in Figure 4.45, draw a small metal rectangle symmetrically placed on the axis of symmetry (the horizontal dashed line). Then change the material to Copper.

Next, select Tools->Add Via->Up One Level and click on the left side of the 2-cell metal rectangle. A small triangle marker appears, which means that a via from GND to the next layer up is created. A via between GND and the patch contacts both GND and the patch. Select Tools->Add Port and click on the left side of the 2 cells, Port 1 is set up and is fed as shown in Figure 4.44. We call this a via port. The height of the via is same as the dielectric thickness of 1 mm. The port is in series with the via.

The position of port is as shown in Figure 4.44 and the size of x direction of patch is 16.5 mm. Figure 4.46 shows the reflection coefficient for switching to this a via port feed line, It is a comfortable -20 dB at 3.97 GHz and the input impedance is 44Ω resistive.

The far field radiation pattern of the patch antenna is shown in Figure 4.47 (Sonnet Lite does not have this feature).

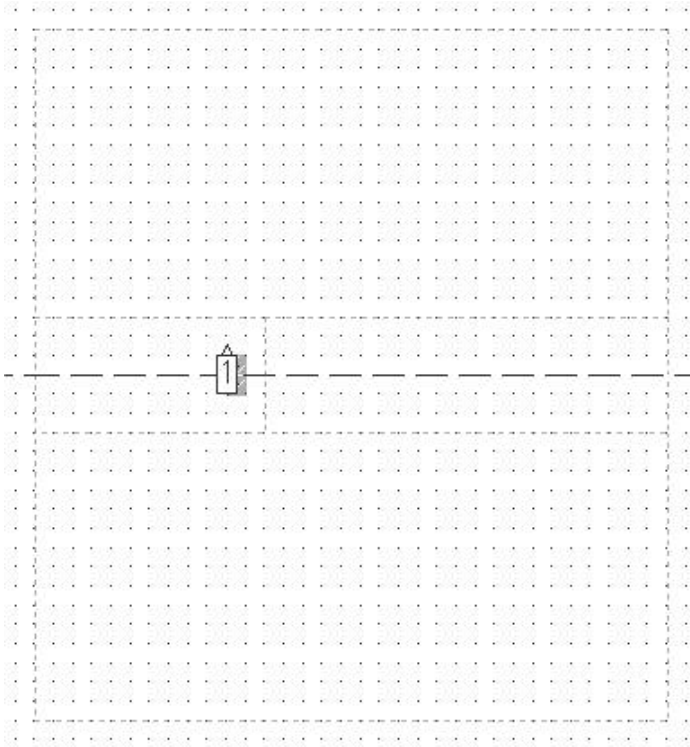


Figure 4.45 Via port as viewed from above the patch antenna level.

The ground plane covers the entire bottom side of the substrate, so there is no radiation downward. The strongest radiation is towards the zenith and its gain is 6.2 dBi (see Chapter 7 for more details).

This plot is based on the coordinate system in Figure 4.48. θ (theta) increases by the angle starting from zenith ($+z$) on the vertical plane and ϕ (phi), or azimuth angle, increases starting from $+x$ on the horizontal plane. Accordingly in Figure 4.47, 0° (degree) Phi means $x-z$ plane and 90° Phi means $y-z$ plane.

4.4 Effects of Dielectric Materials

The dielectric properties of a material refer to how a material changes the electric field passing through it. It can counter the electric field, like a spring, slowing any wave trying to travel through it. It can also transform the electric field into heat, adding loss to any nearby circuitry.

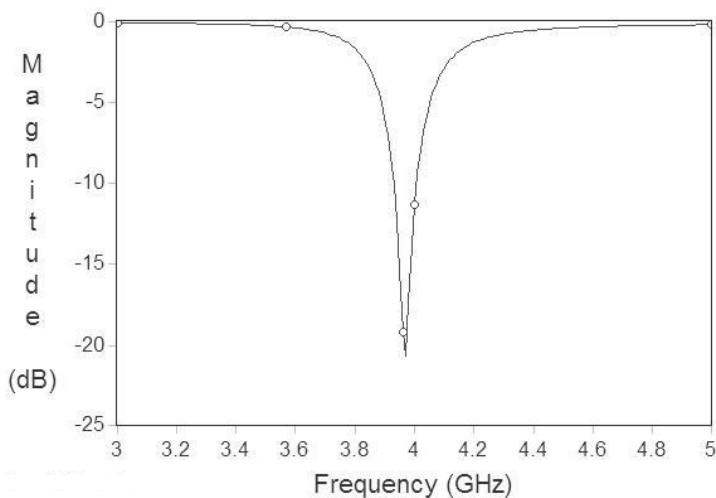


Figure 4.46 Reflection coefficient after switching to a via port feed line, -20 dB at 3.97 GHz.

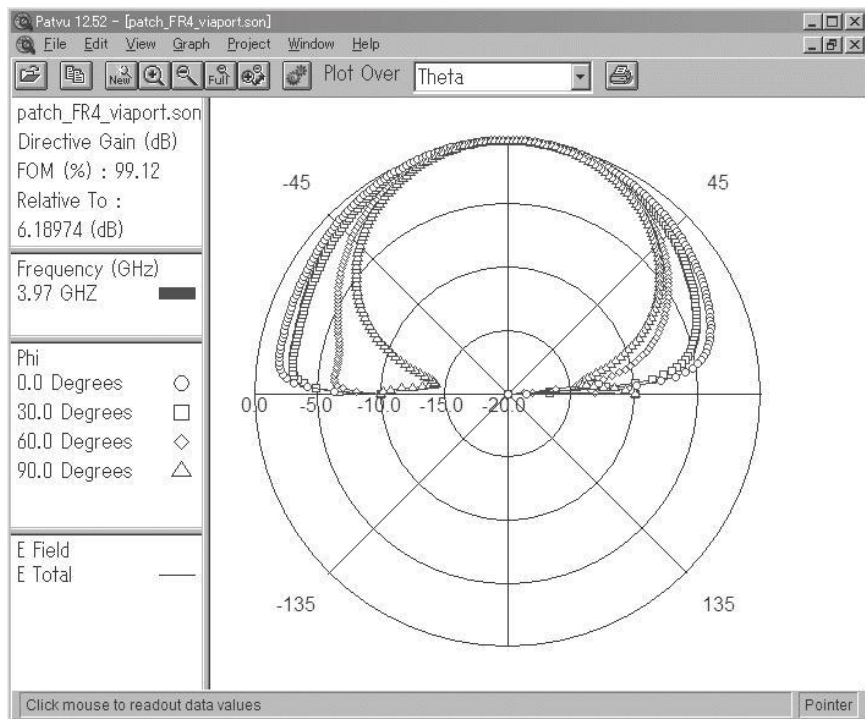


Figure 4.47 Far-field radiation pattern of the patch antenna.

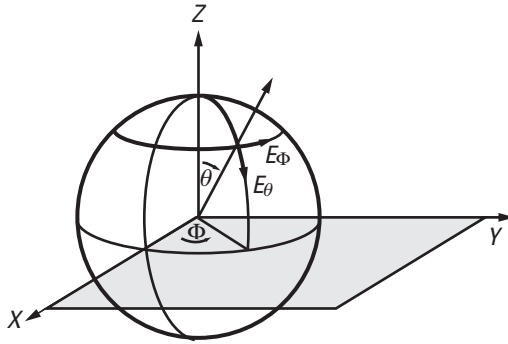


Figure 4.48 Coordinate system used for the radiation pattern shown in Figure 4.47.

4.4.1 Effective Permittivity of Microstrip Lines

Figure 4.49 shows the distribution of electric and magnetic lines of force on a cross section of an MSL. Notice that most of the electromagnetic energy travels in the dielectric and some travels in the air above the substrate. The velocity of a wave in a material is determined by the dielectric constant, or the *permittivity* of the medium. The effective permittivity, or *epsilon effective*, ϵ_{eff} is a weighted average of the dielectric constant of air, 1.0, and that of the substrate. The weighting of the average depends directly on the ratio of the electric field energy in the air to the electric field energy in the substrate.

We should also keep in mind that we usually quote “relative” dielectric constants. That is the dielectric constant is relative to the dielectric constant of free space. So free space is exactly 1.0. Air is actually a tiny bit higher, but it is so close to that of free space, for most applications, they are considered to be the same.

The effective permittivity of the model in Figure 4.40 is shown in Figure 4.50. To get this plot, right-click on DB[S11] in the left frame of display and

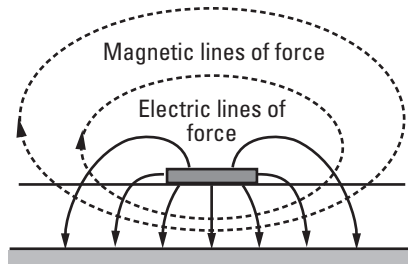


Figure 4.49 Distribution of electric lines of force and magnetic lines of force on a cross section of the MSL.

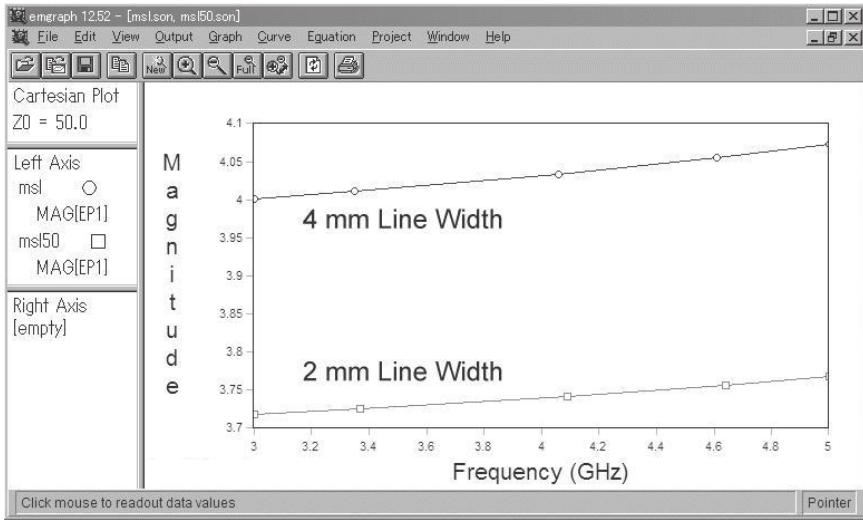


Figure 4.50 Effective permittivity derived from the simulation of the MSL model in Figure 4.40. The line width 4 mm (small circles) and 2 mm (small squares).

select Edit Curve Group, and select Port Eeff from the Data Type drop-down menu. A plot of effective permittivity is then displayed.

The small circle marker is for an MSL 4 mm wide, and its effective permittivity is around 4. The small square marker is for an MSL 2 mm wide and its effective permittivity is slightly less, around 3.7. The narrower line allows more electric field in the air, so the effective dielectric constant is less.

Figure 4.51 shows two parallel plates of area S , separation d . A smaller area, qS , is filled with a dielectric of permittivity ϵ_r . The capacitance C is given in the following expression.

$$C = \frac{\epsilon_0 q S \epsilon_r}{d} + \frac{\epsilon_0 S (1 - q)}{d} = \frac{\epsilon_0 S}{d} \{1 + q(\epsilon_r - 1)\} = \frac{\epsilon_0 S \epsilon_{eff}}{d}$$

where q is the filling factor. This means that the following expression gives the effective permittivity.

$$\epsilon_{eff} = 1 + q(\epsilon_r - 1)$$

For the MSL shown in Figure 4.49, given that the capacitance between a line and its ground is C and when we make the permittivity $\epsilon_r = 1$ everywhere, the capacitance is now C_0 , then the following expression gives us the effective permittivity.

$$\epsilon_{\text{eff}} = \frac{C}{C_0}$$

In the case where most of the electric fields are in the substrate, like a patch antenna, the effective dielectric constant is the same as the relative permittivity (i.e., relative dielectric constant) of the substrate material. On the other hand, in the case where only a portion of the electric field energy is in the substrate, the precise effective permittivity is not easily calculated. In this case, we need electromagnetic field simulation. Sometimes we can use carefully developed closed form equations (like in AppCAD), but these equations, in turn, usually come from detailed EM analysis.

4.4.2 Loss Tangent of Dielectrics

Dielectric material can be represented as an equivalent parallel RC circuit. The capacitor represents a lossless dielectric and the resistor represents dielectric loss. When current flows in this circuit, the relation of current flowing in the capacitor, I_C , and current flowing in the resistor, I_R , is shown in Figure 4.52.

We call δ the dielectric loss angle and $\tan(\delta)$ is called the loss tangent, which we set in Sonnet. The loss tangent is listed on the specification sheets for most substrate materials. Most substrate materials are insulating. However, if a substrate is also a conductor (like silicon), loss increases as well. Conducting substrates are not considered here.

4.5 Effect of Magnetic Materials

In direct analogy with dielectric materials, magnetic materials change the magnetic field passing through them. While everything seems to affect electric fields, only a few special materials affect magnetic fields. Still, it is important to explore.

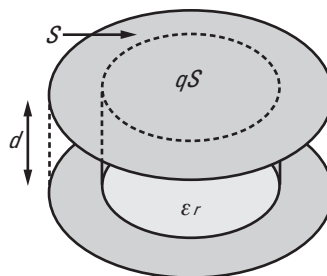


Figure 4.51 Parallel plates filled with a dielectric.

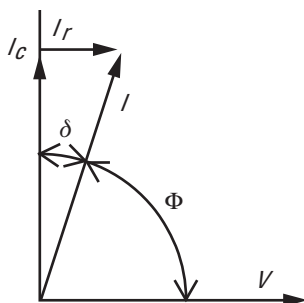


Figure 4.52 Dielectric loss tangent δ and dielectric phase angle Φ .

4.5.1 Characteristics of Magnetic Materials

Insulating materials such as a glass and plastic PET bottles are dielectrics. Human bodies can be modeled as conducting dielectrics. In contrast, ferrites and other magnetic materials can be formed from various mixtures of “soft” magnetic materials. A soft magnetic material does not retain any permanent magnetization. Many different shapes, such as sheets or toroidal cores, can be formed.

The special characteristic for magnetic material is the magnetic permeability, μ . Again, we generally assume that relative numbers are used, so that free space or any nonmagnetic material is 1.0. The more that a material affects magnetic field, the higher μ becomes. Sometimes it is easiest to imagine magnetic field lines more easily “flow” through high μ magnetic material.

The loss in magnetic material is represented by $\tan(\delta_\mu)$, similar to dielectrics, and values are likewise available from manufacturers. The higher that $\tan(\delta_\mu)$ is, the more that magnetic field energy is absorbed. Sometimes a large $\tan(\delta_\mu)$ is desired, in which case the material can be used as a sheet to suppress noise caused by electromagnetic waves.

If you encounter a structure with layers of magnetic materials, you can enter both μ and $\tan(\delta_\mu)$ in the Sonnet Lite dielectric layers menu.



5

Traveling Wave Antennas

5.1 Turning Transmission Lines into Antennas

So far, we have been investigating standing wave antennas, like the dipole. But there are also traveling wave antennas. A traveling wave can travel in free space, or it can travel on a transmission line. In this chapter, we explore how to transition a traveling wave directly from a transmission line to free space without going through a standing wave antenna.

5.1.1 Two Parallel Lines Turn into Antennas

Figure 5.1 shows a Sonnet model of parallel lines 120 mm long, 1 mm wide, and with 10-mm separation in free space. There is a port at the far right and it has an open end at the far left where the electromagnetic wave totally reflects. The simulation space (inside the Box) is 512 mm in both the x and y directions. We have set Free Space for both Box Covers (Top and Bottom), 20 mm above and below the line. Figure 5.2 shows the reflection coefficient of the parallel lines and there are three specific frequencies with near zero reflection. What does it mean?

For example, at 3 GHz the reflection coefficient is under -10 dB. Perhaps power is being lost to radiation. If true, then this transmission line has become an antenna. Figure 5.3 shows the surface current distribution at 3 GHz, and we count three standing wave peaks (including one at the port) and three nulls (including one at the end) along the length of the line.

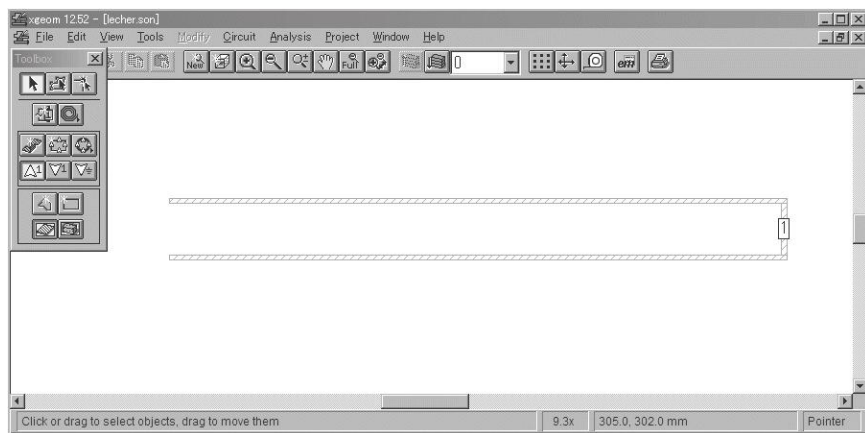


Figure 5.1 Sonnet model of parallel lines. File: lecher.son

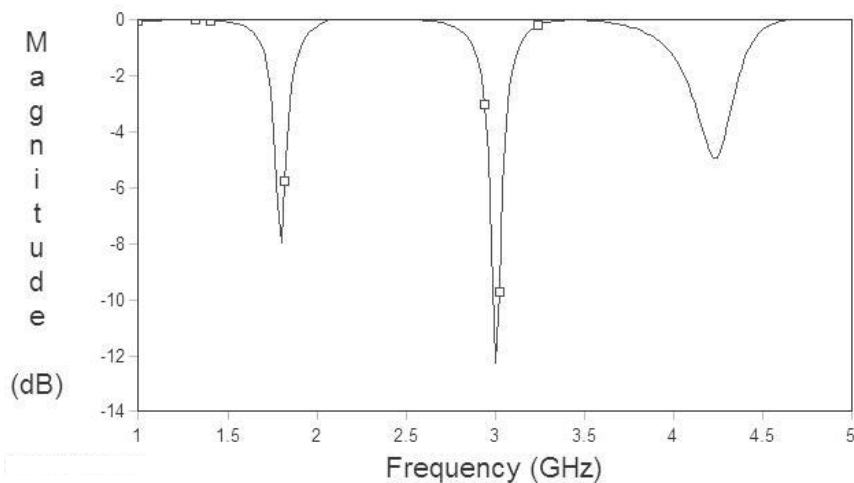


Figure 5.2 Reflection coefficient of parallel lines. There is no reflection at three specific frequencies.

Figure 5.4 shows the electric field distribution around the parallel lines simulated by XFDTD. The current on the parallel lines creates high voltage at the ends of the lines, which in turn creates radiating waves. When we view an animation (not included here), the outer loops of electric field travel into the distance. Thus the transmission line is acting like an antenna.

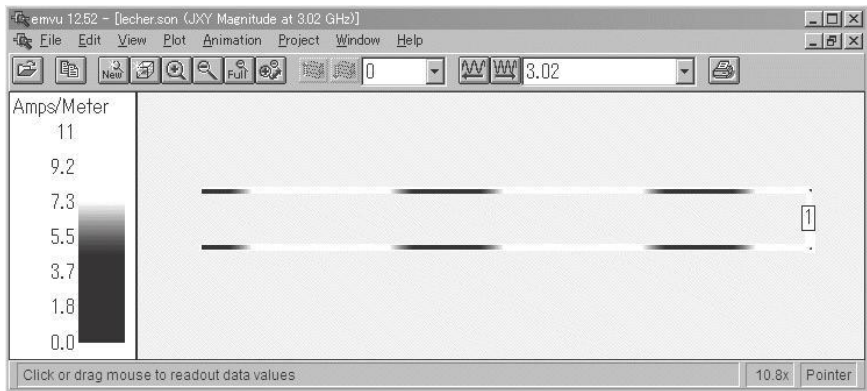


Figure 5.3 Surface current distribution on parallel lines.

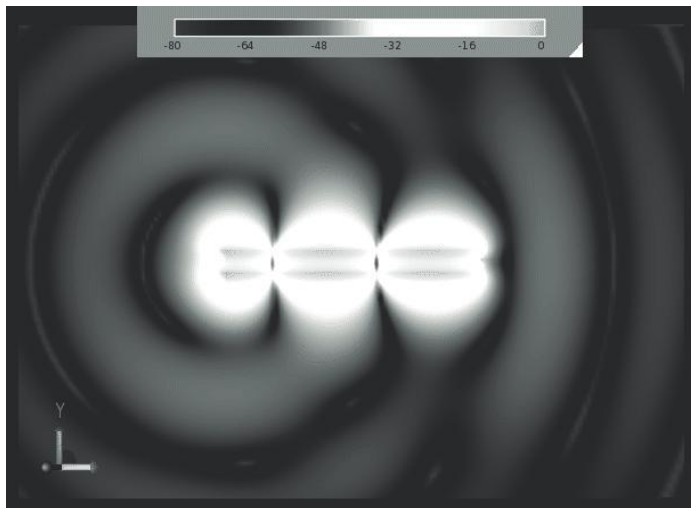


Figure 5.4 Electric field distribution of parallel lines (simulated by XFDTD).

Recalling Chapter 2 where we discussed the electric field (electric lines of force) spreading out from Hertz's dipole, we can imagine that the radiation comes from the ends of lines. One might think that there is no radiation from the middle of the lines. However, the loops of electric field waves in Figure 5.4 depend on the entire line. In this manner, the feed line actually acts as a part of the antenna. We can see this easily when the lines connected to various electronic devices sometime radiate electromagnetic noise (electromagnetic interference or EMI).

5.1.2 The Point of Transition Between Transmission Lines and Antennas

Figure 5.5 is a model of a V-shaped dipole antenna with legs formed by spreading the ends of the parallel lines of the previous example. The loops of the electric field waves do not spread isotropically (equally well in all directions). There is a favored direction, in the same direction as the direction of the traveling wave on the feed line. In addition, they seem to radiate more from a portion of the V.

We see that the radiating electric field depends more on the middle of lines. The main part of the antenna is the V-shaped element. In Figure 5.5, the V angle is set to 90° . If we increase the angle to 180° , it becomes a dipole antenna.

5.1.3 Conditions for a Pure Traveling Wave

The traveling wave launched by the port bounces off the open end of the transmission line/V antenna/dipole. This reflection makes a standing wave antenna, with the loops and nulls in the electric field and current that we have seen above and in many other antennas.

If we were to terminate the ends with a resistance equal to the characteristic impedance of the line, and there is now no reflection, then only a single traveling wave remains. With only one traveling wave, standing waves cannot exist.

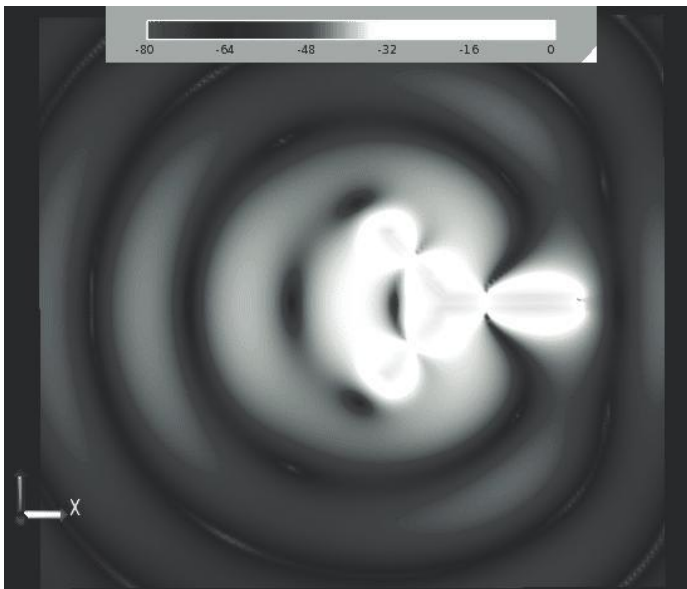


Figure 5.5 Model of V-shaped dipole antenna with legs astride of parallel lines.

Figure 5.6 shows the graph of the characteristic impedance of the parallel lines, when changing Data Type to Port Z0 on the graph shown in Figure 5.2. It covers a range from 260Ω to 310Ω . So, let's simulate with a termination of 280Ω .

You can input the value of resistance in the dialog box that appears after selecting Circuit->Metal Types..., by clicking the Add... button and setting Type to Resistor, and $280\Omega/\text{sq}$ for Rdc (Figure 5.7). Figure 5.8 shows the resulting reflection coefficient with the termination resistor in place. We still see some reflection zeros, which indicate that there is still some reflection at the end of the parallel lines. However, the reflection is much lower than with the original, unterminated, pair of lines (Figure 5.2).

One special form of this antenna uses an earth ground for the second conductor. Because of the earth ground, the antenna is extremely inefficient and thus it is not used for transmitting. However, it is highly directional in the direction that the traveling wave travels. It can be most effective for receiving. If you have sufficient space, it is easily built and works well for receiving signals

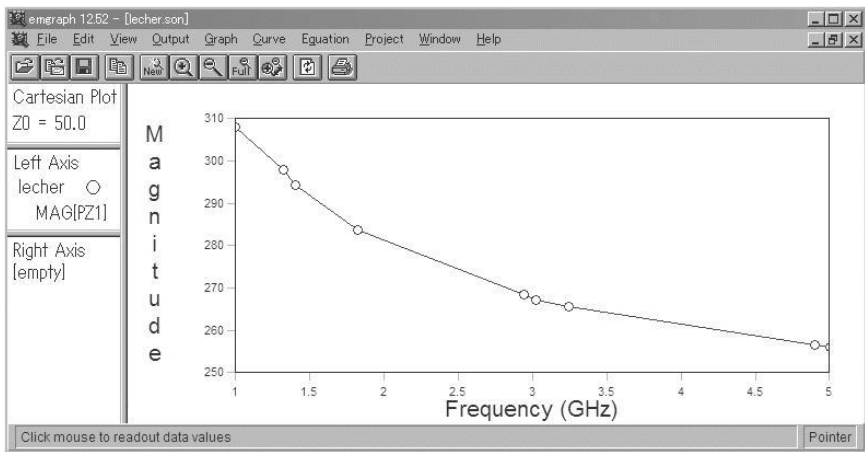


Figure 5.6 Characteristic impedance of the parallel lines.



Figure 5.7 Parallel lines with a 280Ω termination resistor at the end.

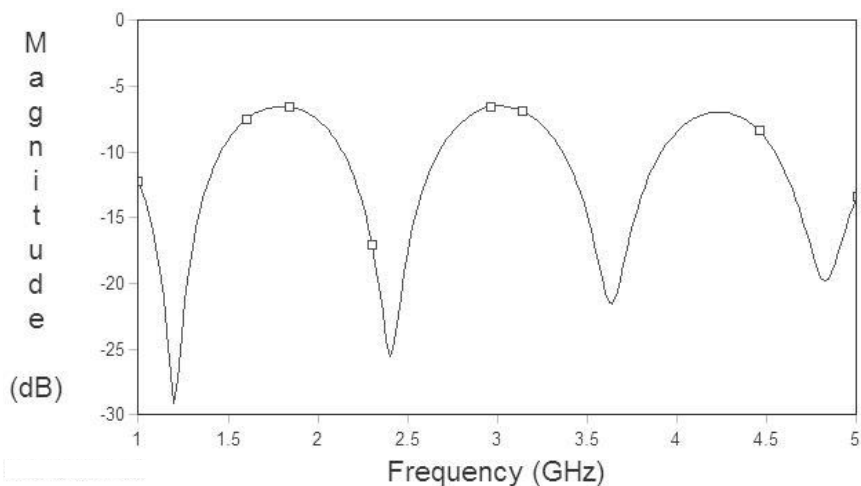


Figure 5.8 Reflection coefficient with the 280Ω termination resistor at the end.

below about 7 MHz. At these low frequencies, there is plenty of signal...and plenty of noise (static crashes). The lack of efficiency suppresses both the signal and the static. The directionality suppresses only the static, leaving an improved signal to noise ratio. In this form, it is called a *Beverage antenna*, named after its inventor.

It seems we get the best radiation when we spread out the two conductors in a transmission line. Figure 5.9 is a rhombic antenna that is created by spreading out the parallel lines halfway between the load and the termination. As

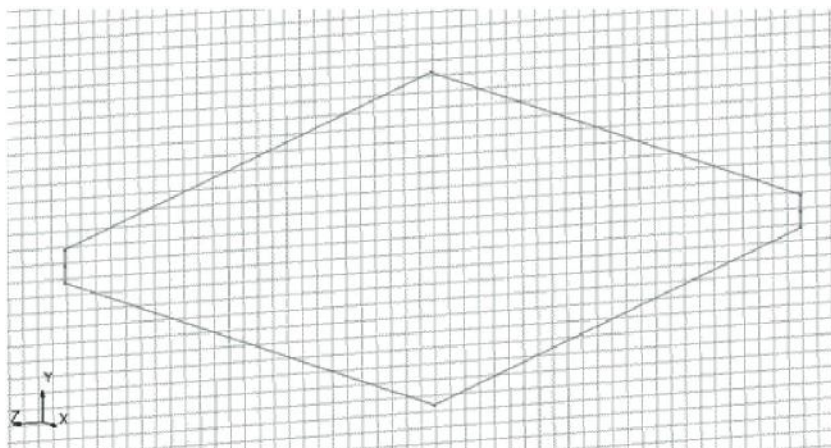


Figure 5.9 Rhombic antenna model.

before, the feed point is at the far right, and the opposite end is terminated with a resistor with the same value as the characteristic impedance of the lines. Because of the termination, this is a traveling wave antenna, which can be thought of as radiating from the midpart of spread-out lines, as suggested in Figure 5.5.

While the radiation pattern does depend on the frequency, it radiates more strongly in the direction of the traveling wave (Figure 5.10). Standing wave antennas tend to be narrowband, and because there is no standing wave in this antenna, it is naturally a broadband antenna. Unlike the Beverage antenna, which uses an earth ground for the second conductor, this antenna is often used for both transmit and receive, as long as the legs are several wavelengths long.

5.2 Antennas That Do Not Resonate

With the antennas that we have discussed above, we are starting to see that they have little if any standing wave patterns in their current distributions. We also see that when the standing wave goes away, we start to get broadband antennas. This could be useful. Let's explore this in more detail.

5.2.1 The Tapered Slot Antenna

Figure 5.11(a) shows how we might radiate right from the edge of a substrate. In the examples above, we gradually spread the two conductors of a transmis-

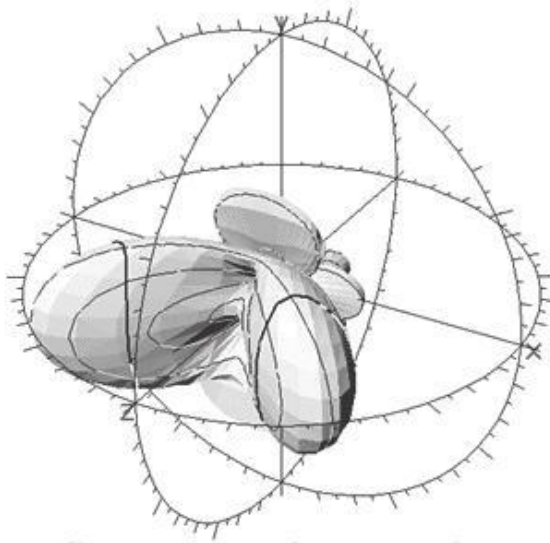


Figure 5.10 Typical radiation pattern of the rhombic antenna.

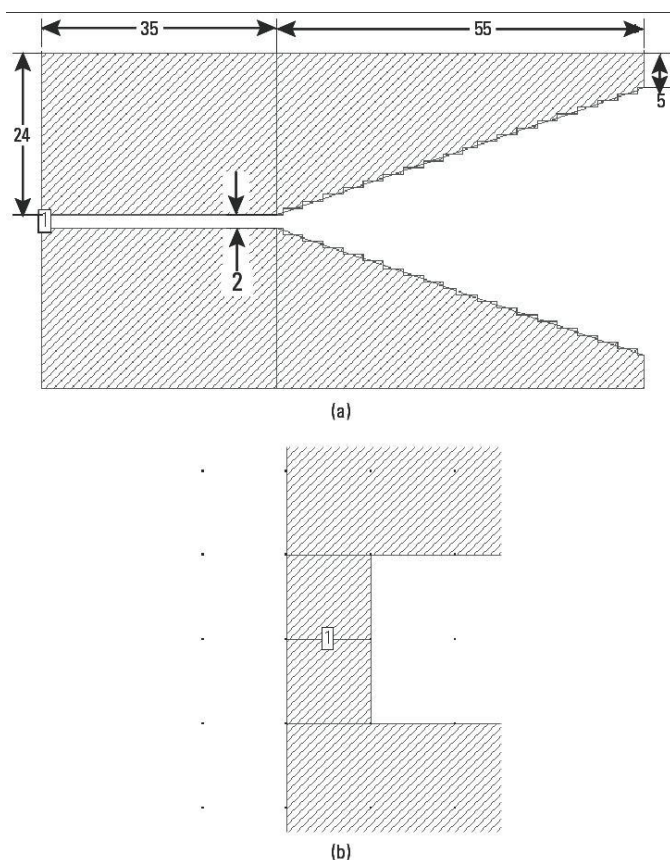


Figure 5.11 (a) The tapered slot antenna. File: TSA.son. (b) Detail of the region near the port.

sion line. Here, we gradually spread the width of a slot line. A slot line is just like the previously described pair of lines, except that the line width is very large. In fact, it is so large that the actual width no longer matters. All the current flows on the inner edges of the slot. By gradually expanding the slot of a slot line, the traveling wave along the slot line can radiate. This is called a tapered slot antenna (TSA).

For this example, the substrate is FR4 with a relative permittivity of 4.9. This is stored in Sonnet Lite libraries, as we found in the previous chapter. The substrate thickness is 1.6 mm, and the feed is located at the far left on the slot line shown in detail in Figure 5.11(b).

Ultra-wideband (UWB), which is used for wireless universal serial bus (USB), needs an ultrawideband antenna that covers 3 to 10 GHz. As for wire

antennas, the rhombic antenna might work, but it is not on a substrate. Many antenna designs have been proposed based on the TSA.

Figure 5.12 shows the reflection coefficient of the TSA in Figure 5.11. As the slot line is fed across a thin slot formed by two metal plates, the characteristic impedance is determined by the width of the slot.

For the model shown in Figure 5.11, the width of the slot is 2 mm, and the characteristic impedance is between 100Ω and 170Ω over the 3- to 10-GHz band. Figure 5.12 shows the reflection coefficient normalized to 130Ω .

The reflection coefficient could be a bit better; however the antenna does realize an ultra-wideband for 3 to 10 GHz. (Sonnet Professional was used because this problem exceeds the 16-MB memory limit of Sonnet Lite.) Figure 5.13 shows the surface current distribution at (a) 3 GHz, (b) 6 GHz, and (c) 10 GHz. At the higher frequencies, many wavelengths are seen along the slot line and tapered region. Even though this is a traveling wave antenna, the low amplitude standing waves are still seen because the antenna is not perfectly matched.

The wavelength at 10 GHz is 30 mm in free space. However, the wavelength we see in Figure 5.13(c) is shorter than that. This is because the relative permittivity of FR4 is 4.9 and, as discussed in the last chapter, this shortens the actual wavelength. With a wave traveling along the tapered slot line, the electric fields fan out in the tapered region, and the loops of electric lines of force spread out and an electromagnetic wave is radiated.

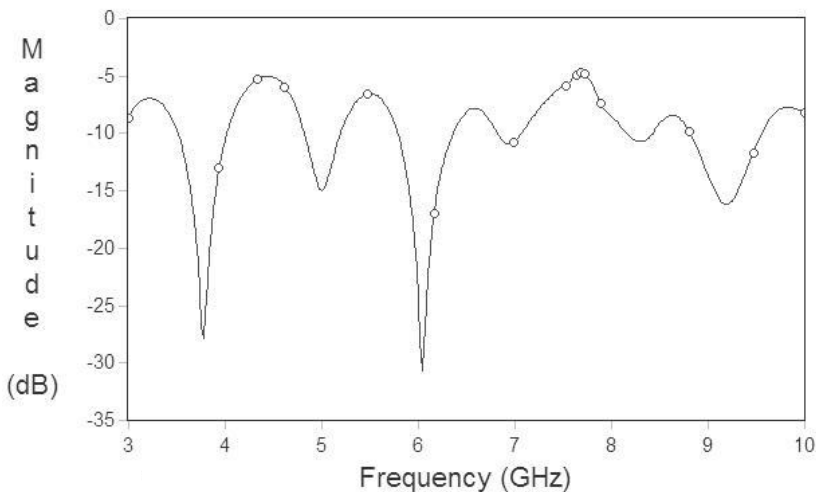
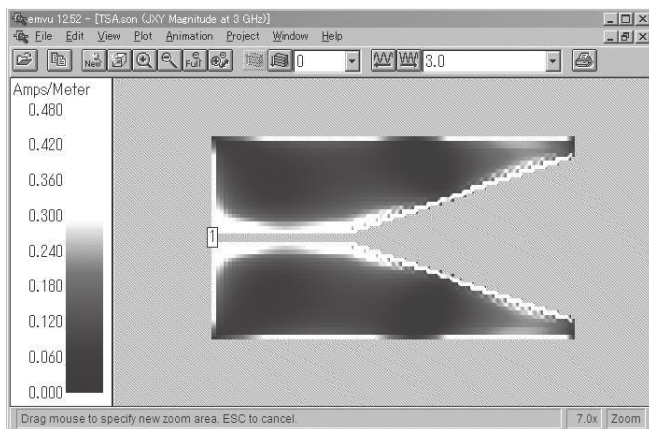
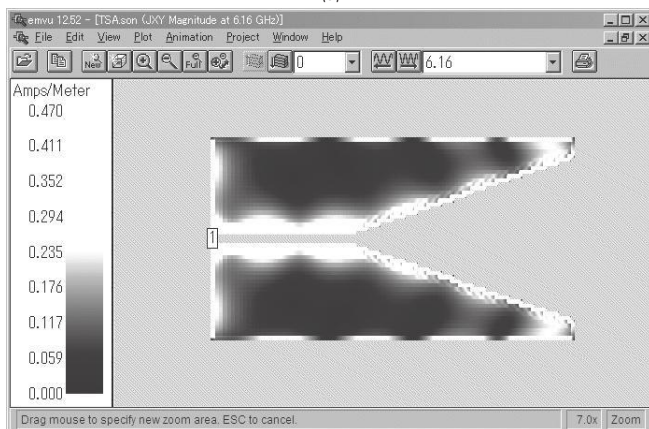


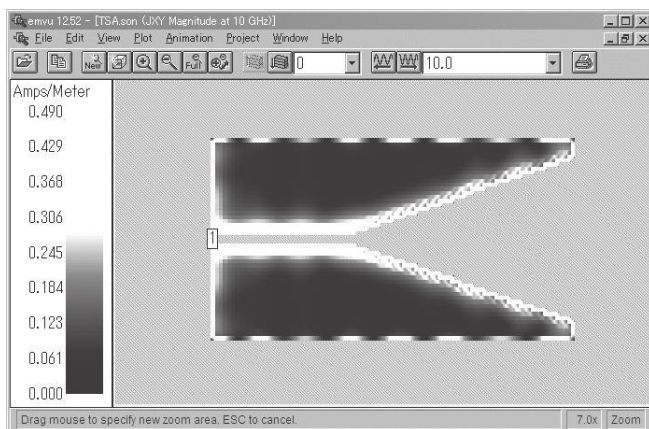
Figure 5.12 Reflection coefficient normalized to 130Ω .



(a)



(b)



(c)

Figure 5.13 (a) Surface current distribution on the TSA at 3 GHz. (b) Surface current distribution on the TSA at 6 GHz. (c) Surface current distribution on the TSA at 10 GHz.

5.2.2 Matching the TSA

The characteristic impedance of this slot is between 100Ω and 170Ω , depending on frequency. We would like to use some kind of matching circuit to match to a 50Ω system. The idea shown in Figure 5.14 is a way to feed the slot line using a microstrip line (MSL) orthogonal to the slot. This method is also used for other slot antennas.

Not all slot antennas are traveling wave antennas. Figure 5.15 is another slot antenna; this one is for 2.4 GHz. The substrate is made of FR4 with a thickness of 1.6 mm, same as above. The current distribution (Figure 5.16) shows two current maximums at either end. There is one current minimum in the center. A current minimum is a voltage maximum. This slot antenna is like a mirror image of a dipole antenna. A dipole antenna has voltage maximums at the ends and a current maximum in the center. Both the dipole and this slot antenna are standing wave antennas.

Because this slot antenna has very high impedance at the center (minimum current, maximum voltage), the feed point is set to be 8 mm to the left of center (Figure 5.15). This should remind you of what we did for the via port of a patch antenna in the previous chapter. This location is where the input impedance is 50Ω . (Sonnet Professional was used for this analysis.)

5.3 Fundamentals of a Bow Tie Antenna

Remember the old bow tie antenna that would come with every TV set? This is just one of a wide class of broadband antennas.

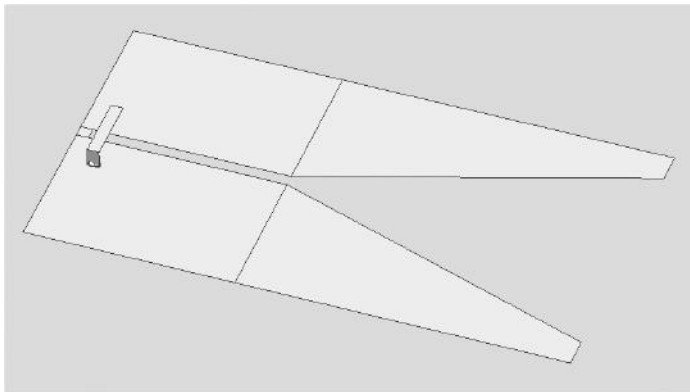


Figure 5.14 Feeding a slot line antenna using an MSL.

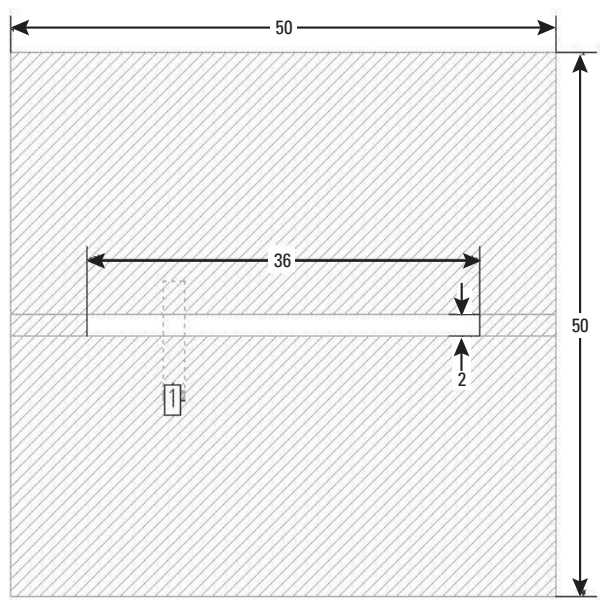


Figure 5.15 Feed system for a slot antenna.

5.3.1 The Biconical Antenna

Figure 5.17 shows a biconical antenna element growing into infinity and its two dimensional equivalent circuit of parallel lines going to semi-infinity. When feeding between the apexes of the two conical conductors, the electric lines of force created between the opposite cones form spherical surfaces. As they move to infinity, a biconical antenna radiates a spherical wave.

The right side of Figure 5.17 shows a parallel transmission line going to semi-infinity; electric fields and magnetic fields travel to the right. This is a two-dimensional version of a biconical antenna. Because there exist only traveling waves for both cases, they have minimal dependence on the frequency.

5.3.2 Finite Length Biconical Antenna

As we cannot build antennas of infinite size, we must truncate the antenna at a reasonable length (Figure 5.18). A finite biconical antenna reflects the traveling wave at the end. The amplitude of the standing wave is smaller if we can minimize the reflected wave.

The right side of Figure 5.18 shows an equivalent transmission line model of the finite biconical antenna. The impedance, determined by the electric and

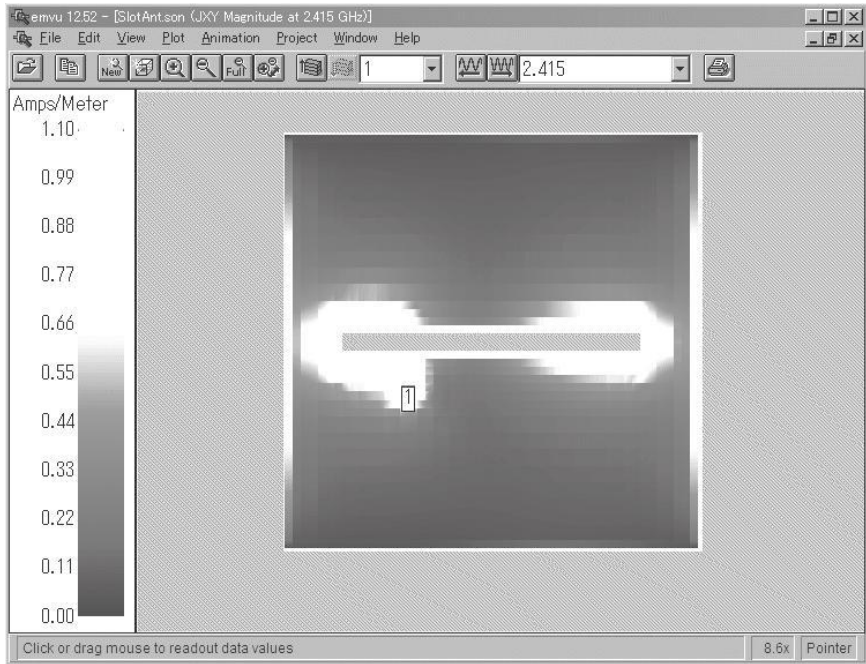


Figure 5.16 Surface current distribution on the slot antenna. File: SlotAnt.son.

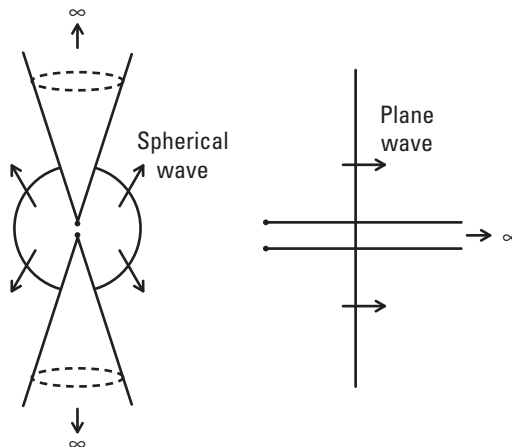


Figure 5.17 The biconical antenna (http://en.wikipedia.org/wiki/Biconical_antenna), and its two-dimensional equivalent circuit.

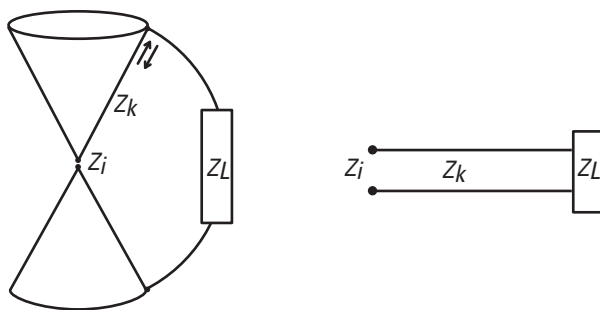


Figure 5.18 Finite biconical antenna and its equivalent transmission line model.

magnetic fields at the end, is Z_L . To minimize reflections, we connect a terminating resistance Z_L at the end. Determining the value of Z_L is a standard electromagnetic problem that we do not detail here. It is sufficient to know that it is easily calculated. It is also interesting to know that the characteristic impedance is not a constant, it is a function of distance from the feed point. Given that the characteristic impedance of a transmission line is Z_k , and the input impedance is Z_i , the condition of $Z_L = Z_k = Z_i$ produces a perfect match.

As the end of element has a transmission line open end discontinuity, there is a reflection. The same goes for the TSA in Section 5.2. In both cases, if you make the antenna longer, say, a few wavelengths, the reflection decreases. The reflection decreases because more power is radiated along the length of the antenna and never gets to the end to be reflected. The same thing happens with the rhombic antenna when the antenna length is a few wavelengths long. If the antenna is long enough and enough power gets radiated before it gets to the end, a termination is not needed.

In any case, longer traveling wave antennas have wider bandwidth and better match. They are not small antennas.

5.3.3 The Impact of Truncating a Traveling Wave Antenna

The engineering trade-off is to reduce the size of the antenna as much as possible, while still meeting the bandwidth requirement. We typically see a critical limit for the element length at around one-quarter of a wavelength long at the lowest desired operating frequency.

The left side of Figure 5.19 shows a planar, two-dimensional finite biconical antenna. The length along the edge of a triangle element is approximately one-half of a wavelength. It looks very much like a half-wavelength dipole with a very wide element.

The left-side antenna in Figure 5.19 is called a bow tie antenna because it resembles a bow tie in shape. The right-side antenna in Figure 5.19 is fed

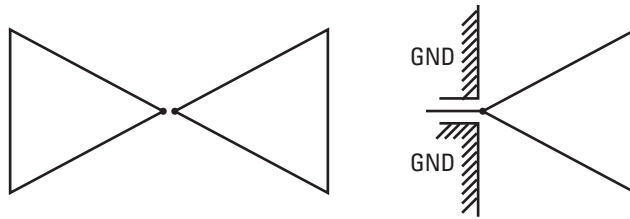


Figure 5.19 Bow tie antenna (left) and triangular antenna (right).

between right-side element and ground. It corresponds to a monopole antenna (Chapter 3).

If our traveling wave is mostly reflected at the end of the element, we would have a standing wave (dipole) antenna rather than a traveling wave antenna. Is it possible that the wide, triangular elements will somehow give us extra bandwidth?

5.3.4 Simulation of a Bow Tie Antenna

Figure 5.20 shows a model of an RFID tag antenna for 950 MHz. The dielectric material is $50\text{-}\mu\text{m}$ thick PET (a kind of plastic) film, and the conductor is $10\text{-}\mu\text{m}$ thick aluminum foil. Figure 5.21 is a graph of its reflection coefficient compared to the dipole antenna in Chapter 3. We do indeed gain significant bandwidth.

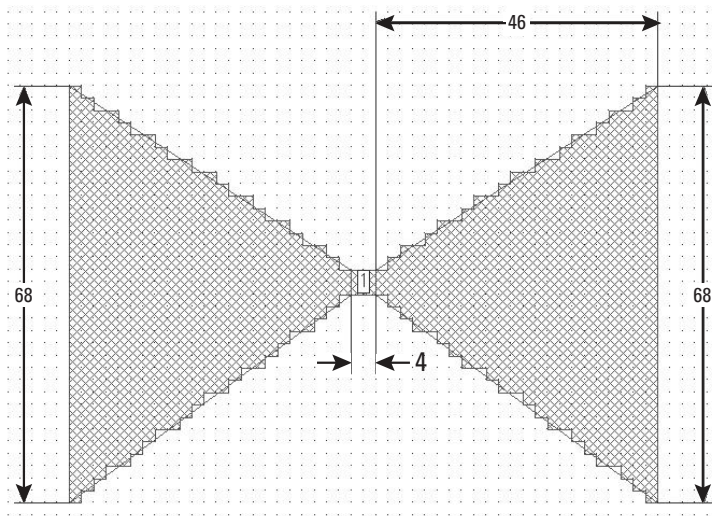


Figure 5.20 Sonnet model of a bow tie antenna.

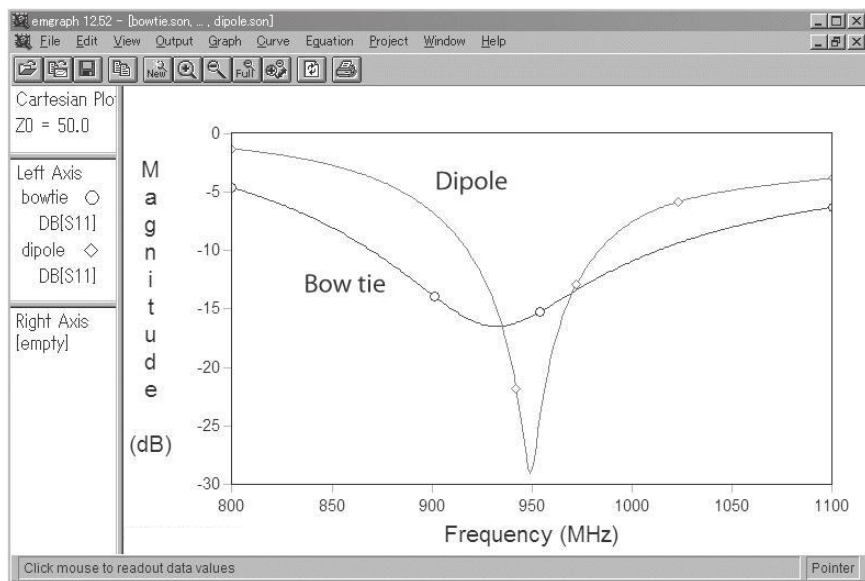


Figure 5.21 Reflection coefficient of the bow tie antenna (circle marker) compared with the dipole antenna (diamonds) of Chapter 3.

5.3.5 Skeleton-Type Bow Tie Antenna

RFID tag antennas are often printed on a PET (plastic) film or on paper with a special conductive ink. Because these RFID tags must be inexpensive, they should be designed to minimize the amount of expensive conductive ink or metal. Figure 5.22 shows the surface current distribution on the bow tie antenna. RF current flows strongest along the edge of each conductor. The narrow neck of each triangle carries the greatest current density.

Try drawing and simulating the antenna in Figure 5.23. By now you should be fairly skilled using Sonnet, so let's introduce an expert level short-cut. With a blank substrate displayed, press CTRL and P at the same time. The cursor will switch into a mode that allows you to draw a polygon. Just click on each point. When finished, click on the last point. Your polygon appears!

The reflection coefficient of the full triangle antenna and the triangle frame antenna described above (Figure 5.24) have nearly the same shape and bandwidth. However, the resonant frequency of the triangle frame antenna is higher, so we need to make it a little longer to resonate at the desired 950 MHz.

Rather than making the antenna arms longer, Figure 5.25 shows how we can connect the ends of the arms together to form a loop element. The apparent resonant frequencies are now almost the same as the full triangle (Figure 5.26).

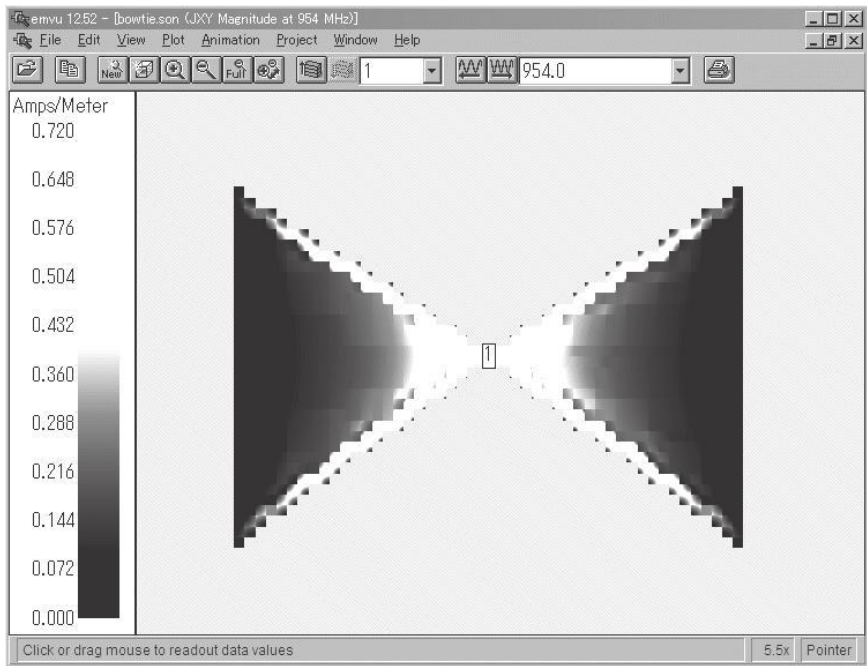


Figure 5.22 Surface current distribution of the bow tie antenna at 954 MHz.

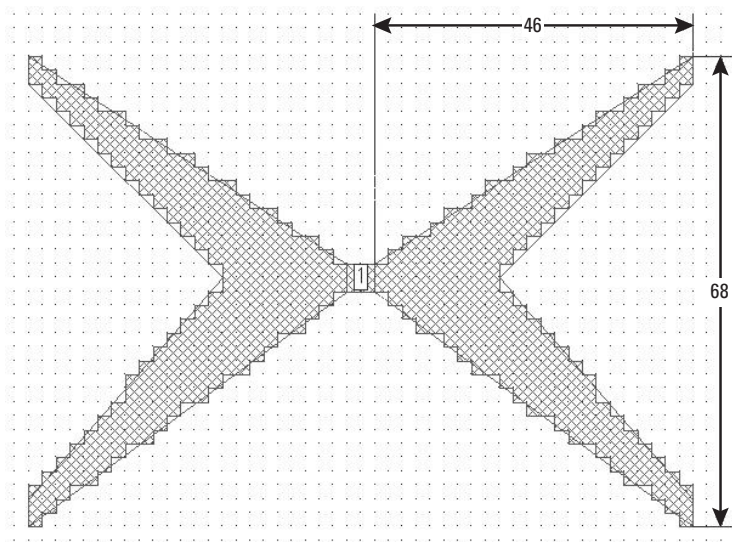


Figure 5.23 Leaving the conductor only where the surface current flows more strongly reduces conductor area. File: bowtie_frame.son.

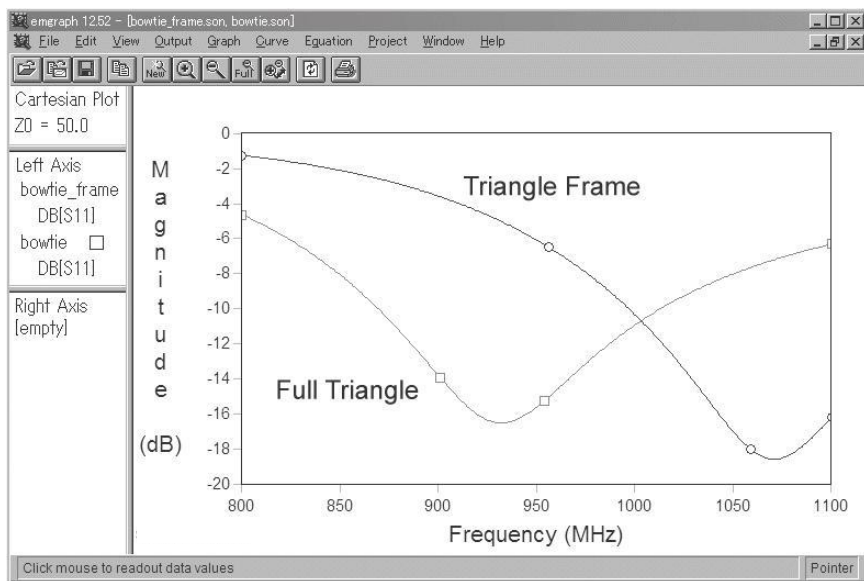


Figure 5.24 Reflection coefficient of the full triangle antenna (square markers) and the triangle frame antenna (circles).

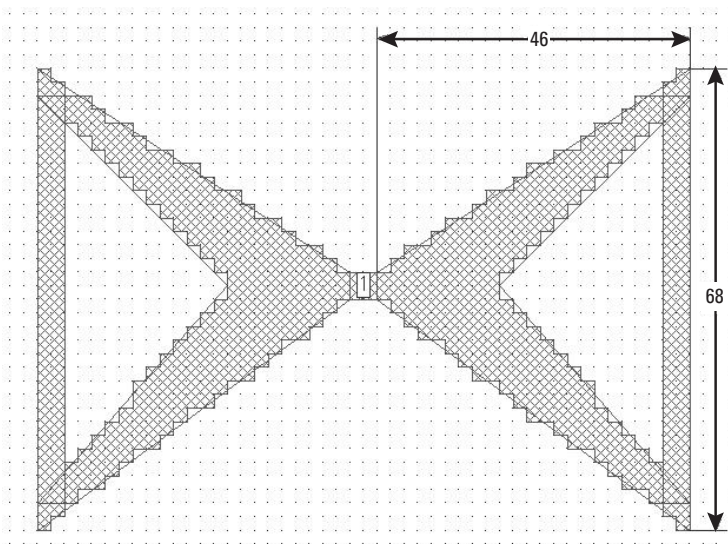


Figure 5.25 Loop element version shorting both ends with straight bars. File: bowtie_frame2.son.

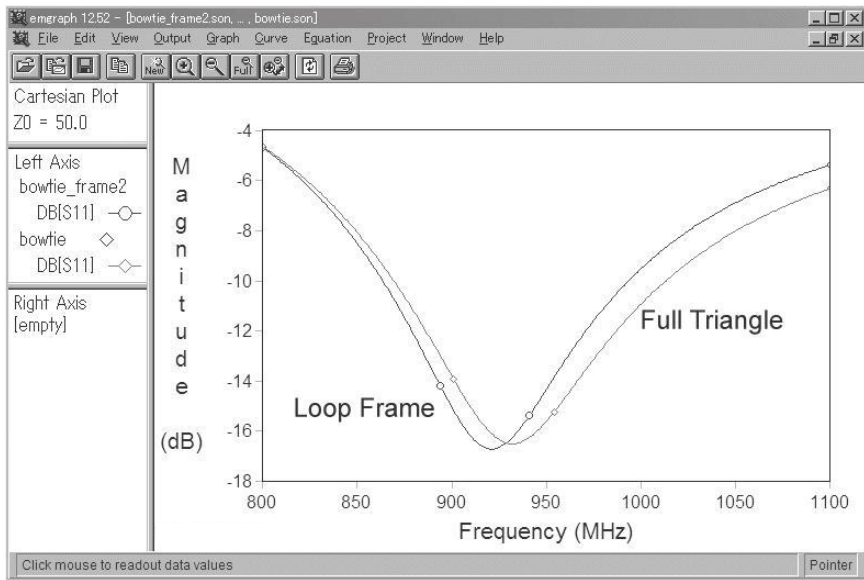


Figure 5.26 Reflection coefficient of the full triangle antenna (diamond markers) and the loop frame (circles) shows almost identical responses.

5.3.6 A Thinner Element Frame

The conducting area can be further reduced by means of thinner elements (Figure 5.27). This time, the apparent resonant frequencies shift a little lower (Figure 5.28). For RFID at 950 MHz, miniaturization can be useful. So we will check the input impedance for confirmation.

Figure 5.29 shows the input impedance for the bow tie antenna R (diamonds), X (triangles), the thinner frame elements R (circle), and X (square). For a bow tie antenna, R is 36Ω and X is zero at 920 MHz. For thinner frame elements, R is also 36Ω and X is zero at 874 MHz.

Reading out the input impedance at the target frequency of 953 MHz (RFID in Japan), $R = 40\Omega$ and $X = 12\Omega$ for bow tie elements, and $R = 45\Omega$ and $X = 38\Omega$ for thinner frame elements. Therefore in both cases the inductive reactance is a bit inductive and the resistance is a little smaller than 50Ω at 953 MHz.

Thinking that the inductive reactance of an antenna is related to the magnetic field component because of extra element length, we can shorten the element length and shape to get zero X at 953 MHz.

This model assumes the RFID tag is for the UHF band (which includes 950 MHz). In fact, the desired input impedance of an RFID IC (integrated circuit) for a tag is typically not 50Ω . Its optimum R and X are different depending

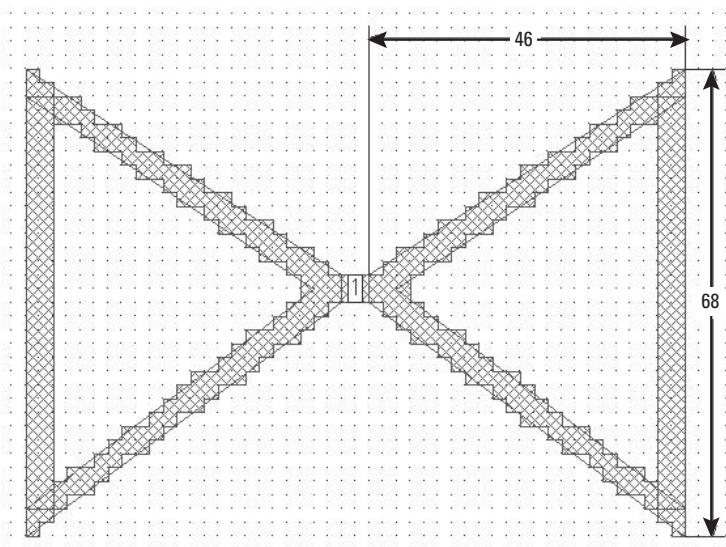


Figure 5.27 Reduced conductor area version with thinner elements. File: bowtie_frame3.son.

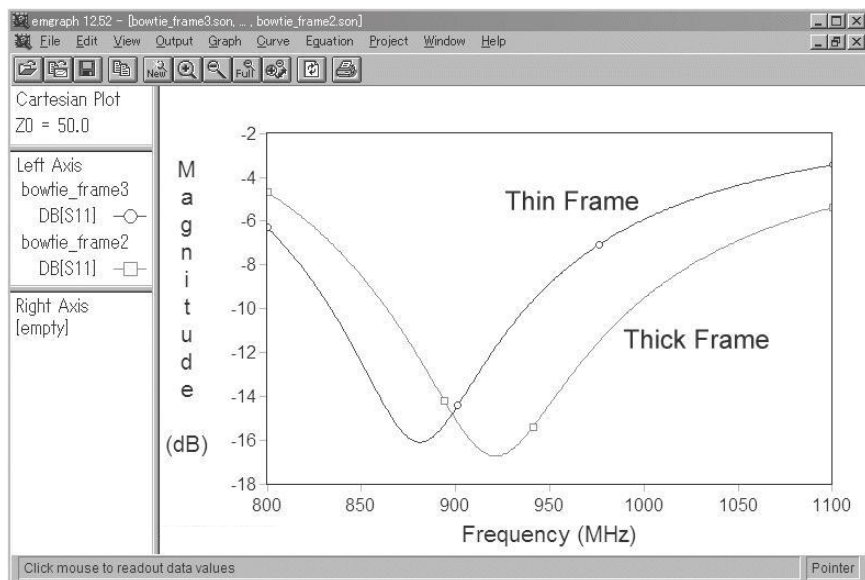


Figure 5.28 Reflection coefficient compared between frame elements (squares) and thinner frame elements (circles).

on the specific IC. In most products the IC is designed with a negative, capacitive, output reactance. We can realize what is called a *conjugate match* by

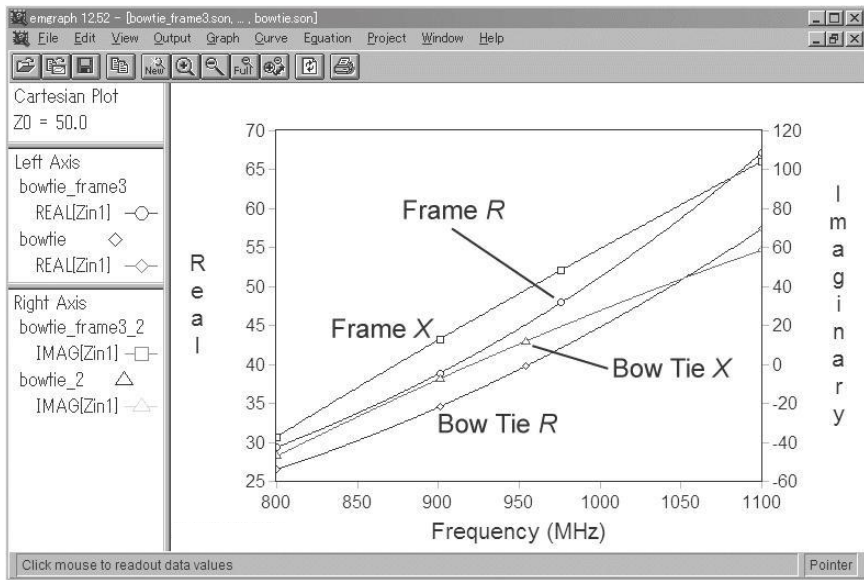


Figure 5.29 Input impedance for the bowtie antenna *R* (diamonds), *X* (triangles) and for the thinner frame elements *R* (circles), and *X* (squares).

designing the antenna with the same amount of positive, inductive reactance. For details see Chapter 8.

Figure 5.30 shows the dimensions for zero reactance. *R* becomes 34Ω at 950 MHz. This is smaller than the bow tie antenna (Figure 5.31).

The reflection coefficient of the full bow tie antenna and the bow tie with thinner frame elements is compared in Figure 5.32. The final version of the thinner frame element antenna has a little narrower band width than the bow tie antenna.

The center frequencies for RFID tag systems are 868 MHz in Europe, 915 MHz in the United States, and 953 MHz in Japan. As we see in the graph in Figure 5.32, the antennas are designed using a little higher frequency. However, if you place these antennas on, for example, a corrugated cardboard box, the dielectric constant of the box shifts the antenna resonance a little lower in frequency because of the wavelength shortening effect of paper (a dielectric material). Depending on the specific situation, it might or might not be a problem in practice.

Figure 5.33 shows the reflection coefficient for a model with 0.5 mm thick paper (relative permittivity is 3.0) just under the antenna. It has shifted a little lower in frequency, covering the three worldwide bands.

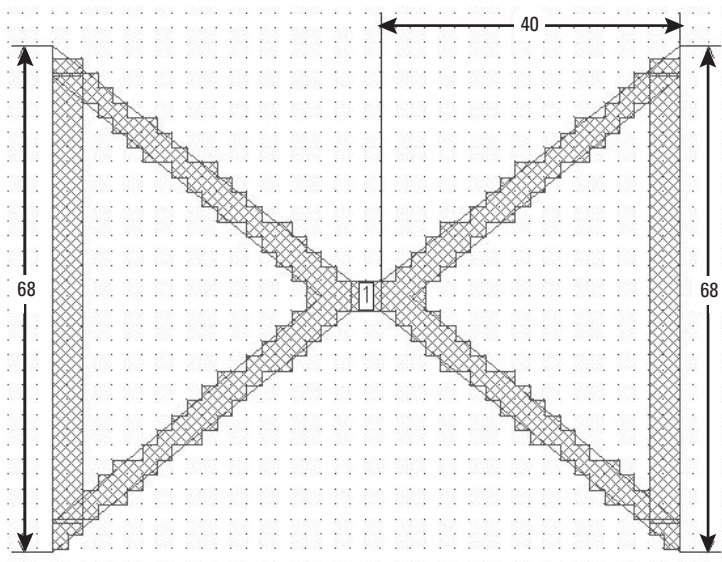


Figure 5.30 Dimensions for zero reactance at 950 MHz. File: bowtie_frame4.son.

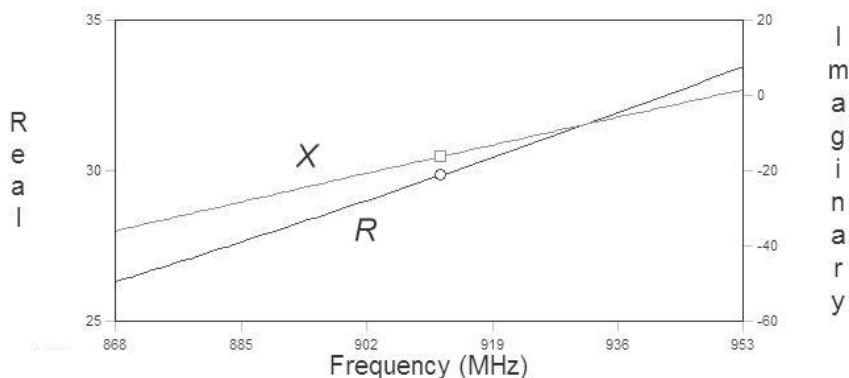


Figure 5.31 R for the thinnest frame bow tie is 34Ω at 953 MHz, just under that of the full bow tie antenna.

5.3.7 Miniaturization Using Triangular Antennas

As shown in Figure 5.19, by feeding between one side element of the bow tie antenna and ground, the antenna corresponds to a monopole antenna and the current on the ground acts like a mirror, simulating the missing element.

Figure 5.34 shows a model of the triangular monopole antenna; the Sonnet box wall is used as a ground plane. After saving the model shown in Figure

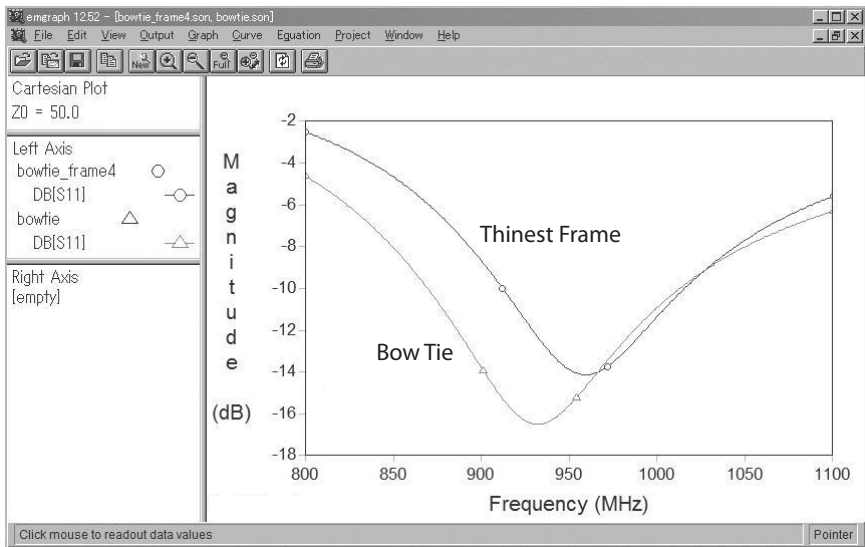


Figure 5.32 Reflection coefficient of the bow tie antenna (triangles) compared with the thinner frame elements (circles).

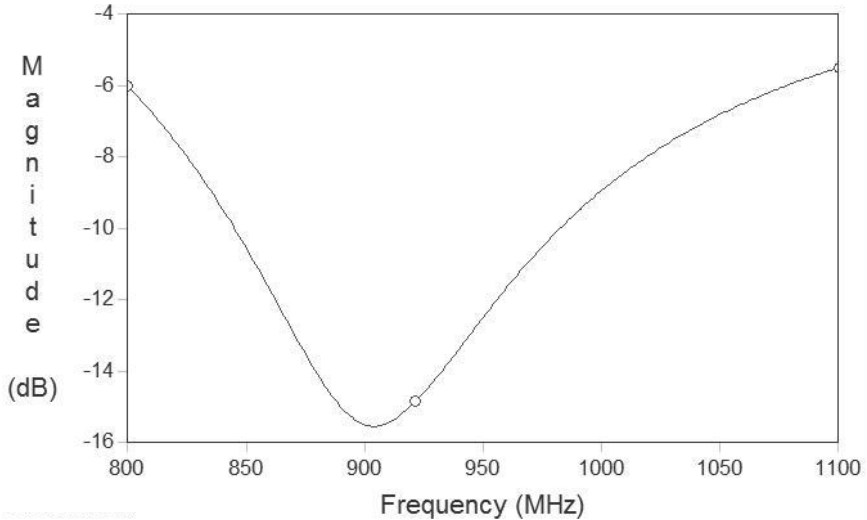


Figure 5.33 Reflection coefficient with 0.5 mm thick paper ($\epsilon_r = 3.0$) just under the antenna.

5.20 under another name, click on the left element structure, and delete it by pressing the Delete key.

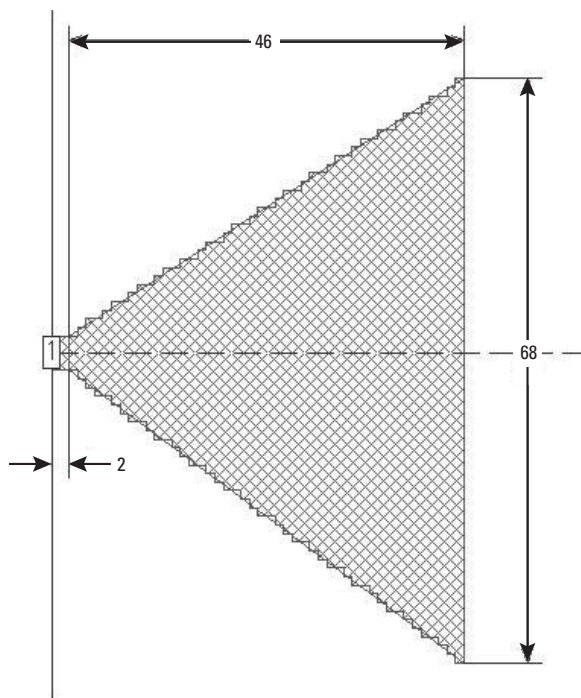


Figure 5.34 Model of the triangular monopole antenna. File: tri.son.

Next, drag the right element next to the left sidewall as shown and set the port.

The horizontal dotted line at the center of the Sonnet box (Sonnet's simulation space) is an axis of symmetry. It is invoked by checking Symmetry in the Circuit->Box... dialog box. When invoked, only the top half of the antenna (above the dashed line) is analyzed with the bottom half automatically included with no extra analysis time. This gives us a faster analysis that uses less memory. To further save memory usage, we increase the cell size to 1×1 mm (Circuit->Box...).

As we learned in Chapter 2, the input impedance of monopole antenna is half the value of a dipole antenna. So the triangular antenna is expected to have an impedance of about 20Ω , which is just half of the 40Ω of the bow tie antenna at 953 MHz.

Figure 5.35 shows the reflection coefficient for this triangular antenna. Normalized to 50Ω , the reflection coefficient at the target frequency of 953 MHz is 7 dB, which is not acceptable. It also seems to be shifting higher in frequency. What is the frequency at which it resonates and X goes to zero?

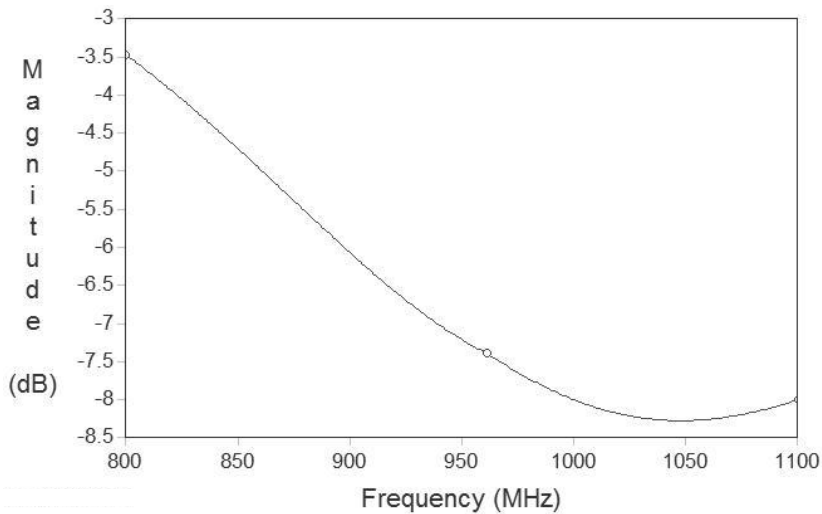


Figure 5.35 Reflection coefficient of the triangular antenna normalized to 50Ω .

The input impedance of the triangular antenna is shown in Figure 5.36. The reactance, X , is almost zero at 953 MHz, and R goes to 20Ω , just one-half of the 40Ω of the bow tie dipole.

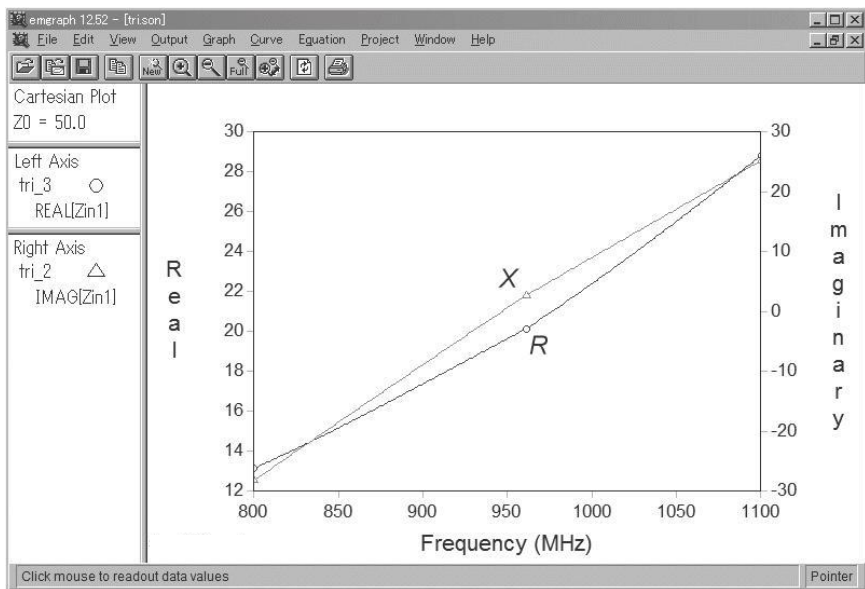


Figure 5.36 Input impedance of the triangular antenna, R (circles), and X (triangles).

As the triangular antenna is shorter than a full element, it can be suitable for miniaturized and built-in antennas. However, the input impedance is lower than a regular dipole. If used in a 50Ω system, it will suffer some mismatch. A matching circuit to transform between 20Ω and 50Ω may be designed using circuit simulators. As we did in Chapter 4, we use S-NAP Design by MEL Inc.

The result of this design is shown in Figure 5.37. As it is realized by a coil of 4 nH in series and a 4 pF capacitor in shunt, we can check it by using a Sonnet Lite Netlist Project. Figure 5.38 shows the input window for the Netlist Project to analyze this matching circuit.

As seen in Figure 5.39, the graph normalized to 50Ω shows the smallest reflection coefficient at 953 MHz, so it will be fine to connect a 50Ω cable directly to the feed point of this matching circuit. Notice that, at least in this case, adding a matching network does decrease the bandwidth.

5.3.8 Flare Angle and Bandwidth

By changing the apex angle (flare angle) at the feedpoint of the triangle element, the electric and magnetic field distribution also changes. In turn, this also changes the input impedance and the bandwidth.

Figure 5.40 shows a triangular antenna with a flare angle of 103° . Figure 5.41 shows one with a 13° flare. The element length (and resonant frequency) can be varied while leaving the flare angle constant.

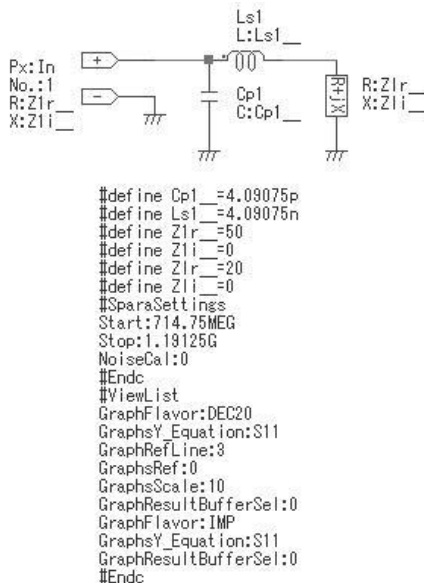


Figure 5.37 Matching circuit created using S-NAP Design.

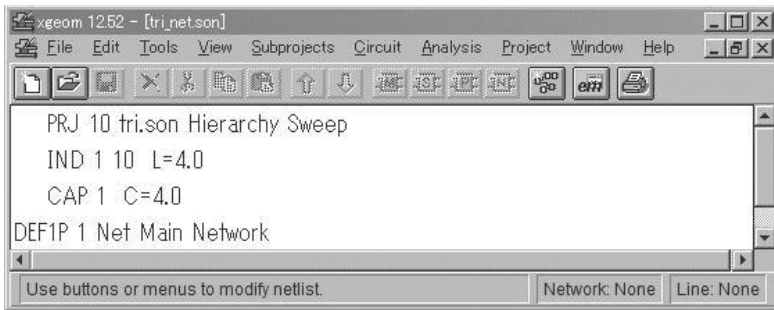


Figure 5.38 Sonnet's Netlist Project window.

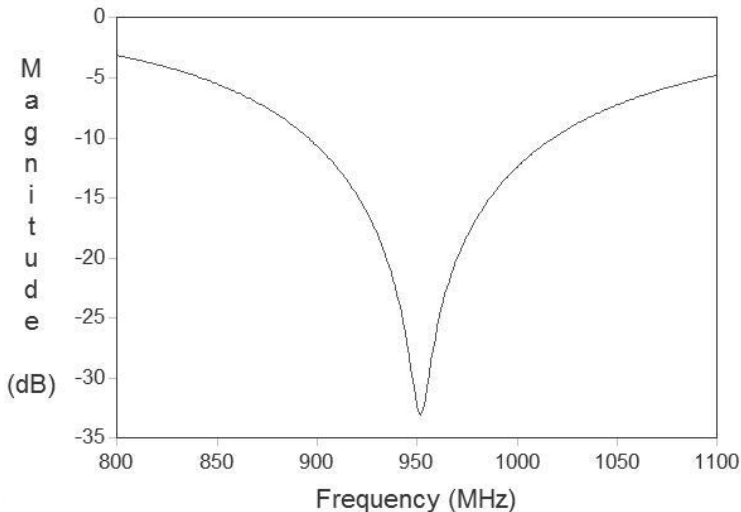


Figure 5.39 Reflection coefficient of the triangular antenna with a matching circuit.

We are covering an extreme range of flare angles. The antenna of Figure 5.41 looks more like a rod rather than triangle. Figure 5.42 shows the reflection coefficient over a range of different flare angles. The antenna with a flare angle of 73° is the best wideband antenna.

5.3.9 A Thin Element Triangular Antenna

Now we will confirm that it is possible to realize a triangular monopole antenna with a thin frame element, just as we did for the bow tie antenna. Figure 5.43 shows a triangular antenna with a thin frame element.

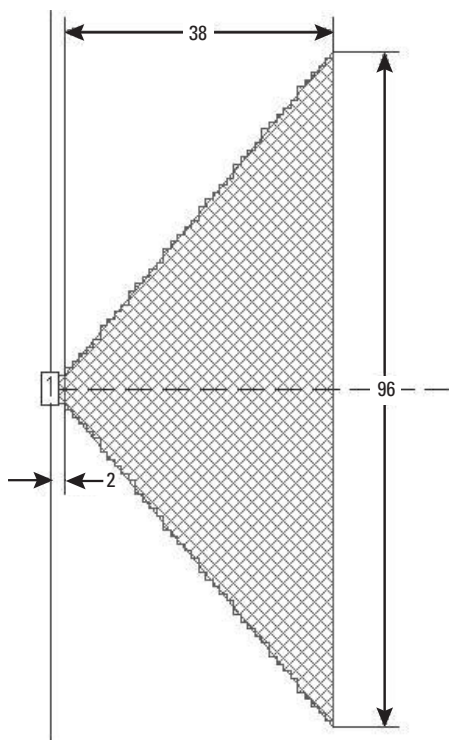


Figure 5.40 A triangular antenna with a flare angle of 103° .

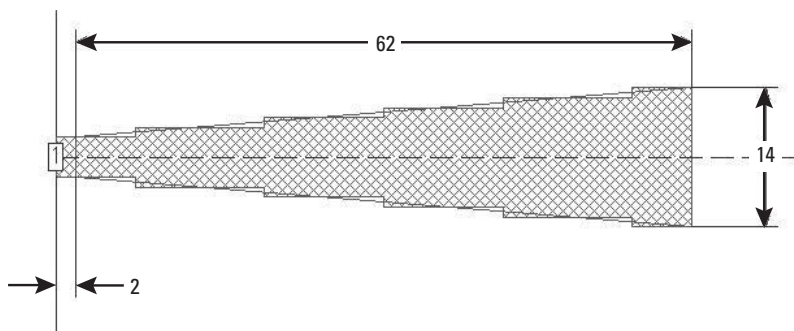


Figure 5.41 A triangular antenna with a flare angle of 13° .

The calculated input impedance is $18.6 + j 5 \Omega$ shown in Figure 5.44. It shows a slight inductive reactance. Setting the resistance to 18.6Ω and reactance to -5Ω by selecting Graph $\rightarrow$ Terminations..., the graph shown in

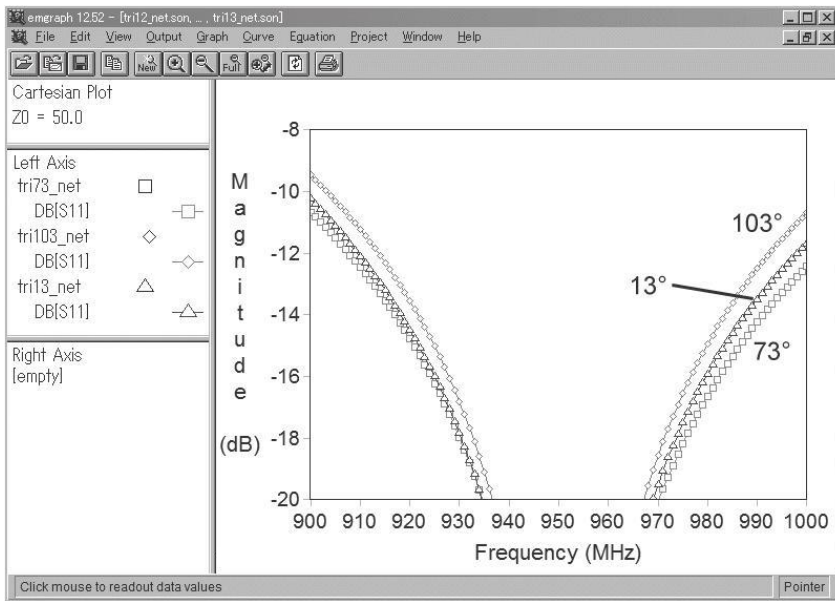


Figure 5.42 Reflection coefficient with flare angle of 73° (squares), 13° (triangles), and 103° (diamonds).

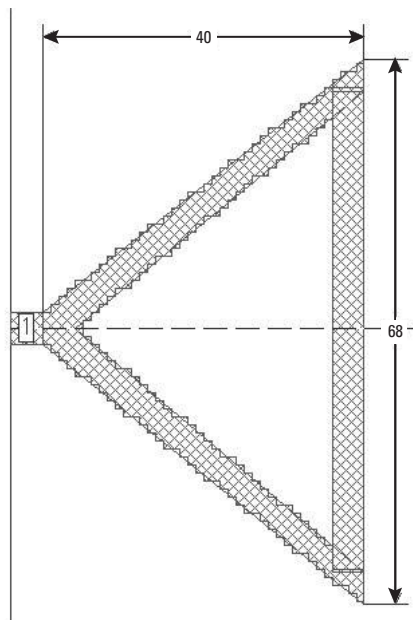


Figure 5.43 A triangular antenna with a thin frame element.

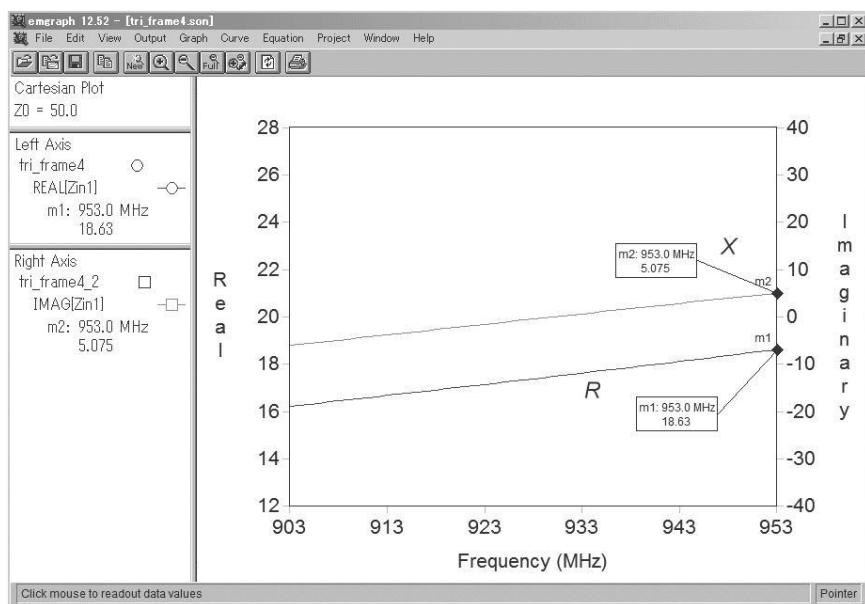


Figure 5.44 Input impedance of the thin frame triangular monopole antenna.

Figure 5.45 appears. This is the reflection coefficient for a conjugate match, mentioned above.

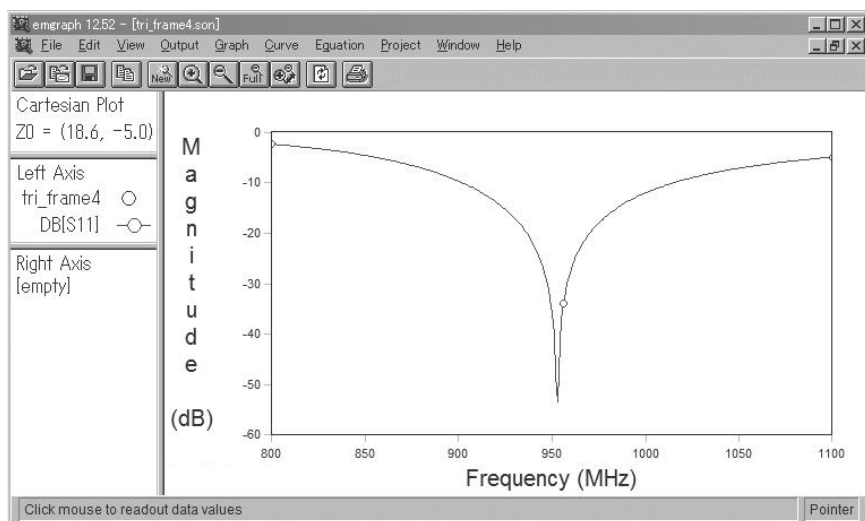


Figure 5.45 Reflection coefficient of the thin element triangular antenna with conjugate matching.

6

Antennas for RFID Systems

6.1 RFID Systems Based on Electromagnetic Induction

One of the big reasons a lot of people are getting into antenna design for the first time is radio frequency identification (RFID). With RFID proliferating everywhere, in the entire supply chain, in price tags, in credit cards, in subway passes, in automobile toll collection, and so on, someone has to design the RFID tag antennas. It might as well be us. Most of what we have already learned is directly applicable. Let's start with antennas that are really inductors.

6.1.1 Faraday's Law of Electromagnetic Induction

Michael Faraday (1791–1867) was a physicist and chemist in England. He discovered the law of electromagnetic induction in 1831, the same year as Clerk Maxwell's birth. He also made many discoveries in the field of chemistry. The unit F (Farad) of capacitance is named after him.

Faraday knew Oersted and Ampere had discovered that magnetism is generated by electrical current. He wondered if it might be possible to do the reverse, to generate electrical current from magnetism.

If the magnetic flux, Φ , (think of the iron filings we saw in grade school science class) increases as shown in Figure 6.1, electromotive force (think voltage), E , is generated. If there is a conductor appropriately placed, then the electromotive force can cause current flow. This changing current flow can in turn modify the magnetic flux. Figure 6.2 is a simplified schematic diagram of the experimental apparatus that Faraday used when he discovered electromagnetic induction.

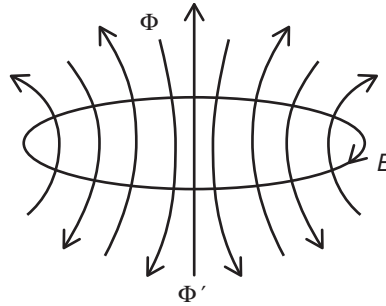


Figure 6.1 Faraday's law of electromagnetic induction.

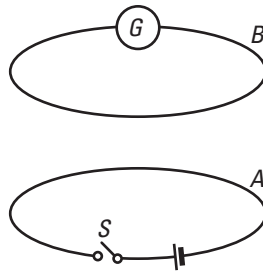


Figure 6.2 Simplified schematic of Faraday's experiment.

When closing switch S and applying current to A , galvanometer G (ammeter) swings a little at that same instant. But after a short period of time, G returns to zero and no longer moves. Next we open switch S , and G again swings at that same instant, but, Faraday found, the current now flows in the *opposite* direction.

The electric current source back then was the Galvanic piles that Volta had developed, which is the ancestor of the battery as we know it today. Galvanic piles generate direct current. What is important here is his discovery that a changing current in coil A creates a changing magnetic flux that passes through both coil A and then into B . The changing flux in B then generates an electric current in B . The important point is that changing currents and fluxes are required. Steady, DC, currents, and magnetic fluxes generate no induction.

6.1.2 Self-Inductance of a Coil

We can see that when we apply current to a loop of wire, magnetic flux appears in the space around it. The total amount of magnetic flux that intersects the

loop of wire is represented as Φ . However, the total amount of flux depends on the shape and size of the loop, among other things. In addition, for any given loop of wire, if we double the current, we also double the flux. Given this, we can infer that there is a relationship between the electric current, I , and the total flux, Φ , as in the following expression.

$$\Phi = LI$$

The proportionality constant L is called self-inductance of this coil. We can also write the following.

$$L = \Phi/I$$

Self-inductance is defined as the total magnetic flux generated by a unit current. When measuring magnetic flux by Wb (Weber) and electric current by A (Ampere), the unit of inductance is Wb/A, which is called H (Henry), named after Joseph Henry (1797–1878), a physicist in the United States who discovered the law of electromagnetic induction independently of Faraday.

Figure 6.3 shows models with an increasing number of turns as analyzed by Sonnet Lite. These models use a SPICE lumped model extraction feature to find the self-inductance, L , of each coil. Notice that there is no such thing as a perfect inductor. Elements such as stray fringing fields make things a bit more complicated by adding some resistance and capacitance, too.

The port for multiturn coils is attached in a short line that passes over the multiple turns and is connected between the vias (the small triangles). This is an application of an “internal” port.

The SPICE equivalent circuit of the one-turn coil has only one L and one R . Multiturn coils generates slight parasitic capacitances between lines. The

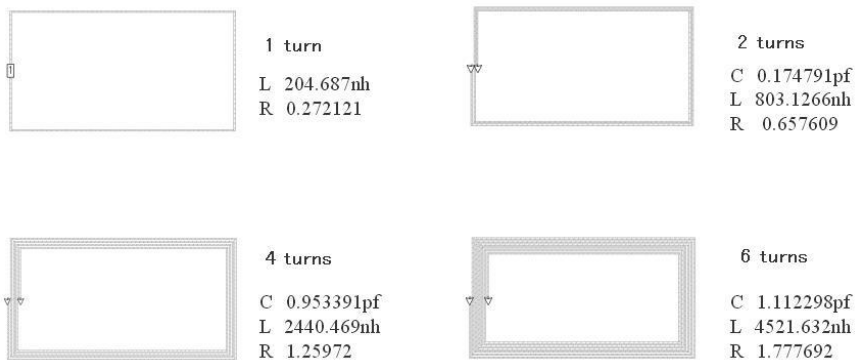


Figure 6.3 Relationship between number of turns and self-inductance L .

resistance of the conductor used for the coil, like copper, determines the value of R , which varies with the number of turns and the size of the conductor.

6.1.3 What Is a Mutual Inductance?

RFID at 13.56 MHz includes the antenna of the reader-writer coil (to be held near the tag) and the tag coil. The two interact via magnetic energy. Thus, strictly speaking, they are not antennas because antennas interact via electro-magnetic energy. However, they are frequently referred to as antennas anyway.

With multiple tags, and defining the current in each coil as $I_1, I_2, I_3, \dots$ and the mutual inductance (i.e., current in one coil causing flux in another coil, which induces voltage in the second coil) is as follows.

First, the magnetic flux coupled into coil 1:

Magnetic flux by I_1 in coil 1 causing flux in coil 1: $\Phi_{11} = L_1 I_1$

Magnetic flux by I_2 in coil 2 causing flux in coil 1: $\Phi_{12} = M_{12} I_2$

Magnetic flux by I_3 in coil 3 causing flux in coil 1: $\Phi_{13} = M_{13} I_3$

and so forth.

Similarly, the magnetic flux that couples into coil 2,

Magnetic flux by I_1 in coil 1 causing flux in coil 2: $= M_{21} I_1$

Magnetic flux by I_2 in coil 2 causing flux in coil 2: $= L_2 I_2$

Magnetic flux by I_3 in coil 3 causing flux in coil 2: $= M_{23} I_3$

and so forth.

Current in any coil generating magnetic flux in any coil is similarly possible. Here the coefficients such as L_1, L_2, M_{12} , and so forth are the ratio of the magnetic flux to the current causing that flux. They are determined by the specific coil shapes, locations, orientations, and other factors.

For instance, M_{12} is called the mutual inductance between coil 1 and coil 2. Thus mutual inductance is the total flux generated in one coil by current in another coil.

6.1.4 Coupling Coefficient Between Reader–Writer’s Coil and the Tag Coil

As shown in Figure 6.4, we have a tag coil and a reader coil in free space. Given that the self-inductance is L_1 and L_2 , and mutual inductance is M_{12} , when electrical current I_1 and I_2 flow in these coils, the magnetic energy W is represented by the following expression summed over all possible pairs of coils:

$$W = \frac{1}{2} (\text{current in one coil}) \times (\text{magnetic flux coupling to another coil})$$

From the above expression, we have for two coils:

$$W = \frac{1}{2} (L_1 I_1^2 + L_2 I_2^2 + 2M_{12} I_1 I_2) = \frac{1}{2} \left\{ L_1 \left(I_1 + \frac{M_{12}}{L_1} I_2 \right)^2 + \left(L_2 - \frac{M_{12}^2}{L_1} \right) I_2^2 \right\}$$

Because we cannot have negative energy, this expression is not allowed to be negative. This means $L_1 \geq 0$ and $L_2 - \frac{M_{12}^2}{L_1} \geq 0$, which means $L_1 L_2 - M_{12}^2 \geq 0$ is required for a physical, passive inductor given that L_1 and L_2 also must be positive.

Notice that mutual inductance M_{12} can be negative as long as it is less than the product of L_1 and L_2 . If we have two coils that have a positive mutual inductance, we can easily create a negative mutual inductance by reversing the leads on one of the inductors. Alternatively, we can just wind one of the coils in the opposite direction.

In Figure 6.4, some of the magnetic flux generated by I_2 couples through loop 1 and induces I_1 in the same (positive) direction, so M_{12} is positive. We get to arbitrarily decide which direction is positive. If we change definition of positive I_1 or I_2 (not both), M_{12} is negative. We see that when we define the direction of positive current in both coils to be the same, the mutual inductance is positive. When opposite, it is negative.

All magnetic flux generated by current on coil 1 passes through coil 1. All magnetic flux generated by current on coil 2 passes through coil 2. In Figure 6.4, all the magnetic flux generated by coil 1 also passes through coil 2. All the magnetic flux generated by coil 2 also passes through coil 1. In other words, the two coils are perfectly coupled, without flux leakage.

Any magnetic flux generated by a coil that does not pass through another coil is called *leakage flux*. In this case:

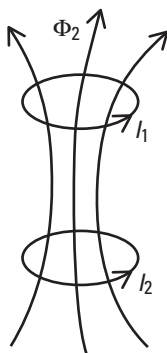


Figure 6.4 Two coils in free space.

$$L_1 L_2 > M_{12}^2$$

Here, the smaller M_{12}^2 is, the more leakage flux exists between the two inductors. This prompts us to define the coupling factor, k .

$$k^2 = \frac{M_{12}^2}{L_1 L_2}$$

In other words, there is no leakage flux when the coupling coefficient k is 1, and we call it *tight coupling* when k is close to 1.

6.1.5 Finding the Coupling Coefficient k Using Sonnet Lite

Figure 6.5 is a model with two coils separated by 10 mm. The cell size is 1×1 mm, and the 70×40 mm rectangle loop whose line width is 1 mm is fed by two parallel lines. In this case, when modeling two floating coils, a common ground does not exist. A common ground is needed to generate a correct SPICE RLC model, so the Box wall ports are needed. In order to exclude the effects from port connecting lines, we set a reference plane at the end of each coil.

To obtain an equivalent circuit (subcircuit) SPICE model at 13.56 MHz as shown in Figure 6.6, we set two analysis frequencies, 13 MHz and 14 MHz, set in Linear Frequency Sweep.

Next, in the dialog box displayed after selecting Analysis->Output Files, click in the upper right on PI Model and change the value of Lmax in middle column to 1,000, and click OK (Figure 6.7). The default (initial) value of Lmax

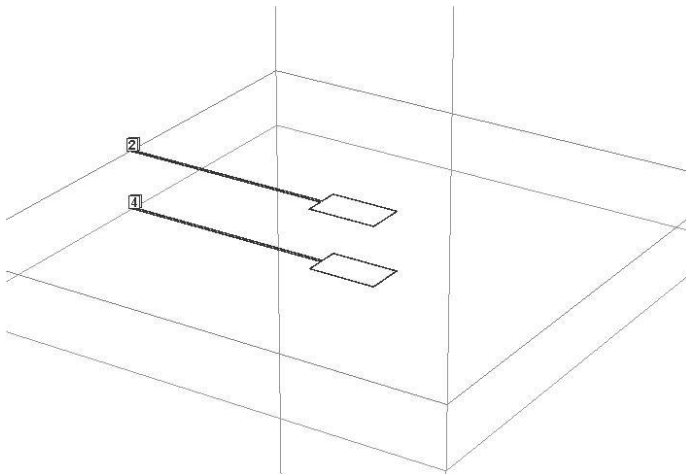


Figure 6.5 Sonnet model, two coils separated by 10 mm.

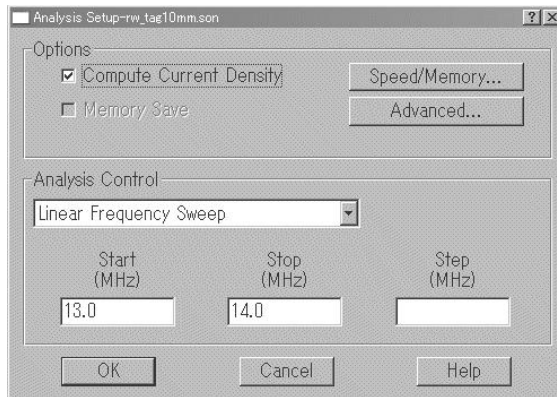


Figure 6.6 Two frequencies, 13 MHz and 14 MHz, are set in Linear Frequency Sweep.

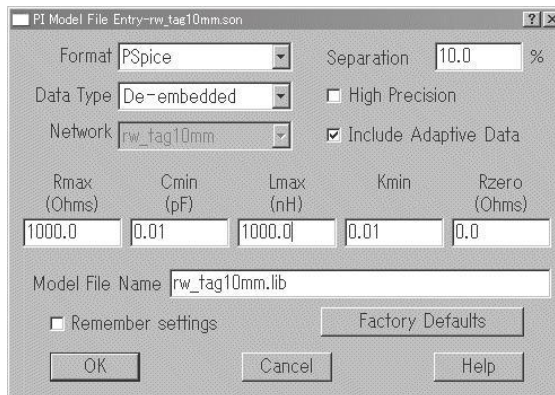


Figure 6.7 The default value of Lmax should be changed to 1,000.

is 100 nH. If unchanged and the coil model requires more than 100 nH, model synthesis will be unsuccessful.

By selecting Project->Analyze and running a simulation, a text file with extension “.lib” is generated. Figure 6.8 shows the resulting subcircuit in PSPICE format, a model for the two coils. L is 197 nH, and the coupling coefficient, k , is 0.22.

6.1.6 13.56-MHz Antenna (Coil)

Figure 6.9 shows a 6-turn coil modeled by Sonnet Lite. The cell size is 0.5 mm square and the simulation space (Box) x and y are set to 256 mm. Figure 6.10 is

```

* Spice Data
* Limits: C>0.01pF L<1000.0nH R<1000.0ohms K>0.01
* Analysis frequencies: 13.0, 14.0 MHz
.subckt rw_tag10mm_0 1 2 3 4 GND
C_C1 1 GND 0.562831pf
C_C2 1 4 0.840164pf
C_C3 2 GND 0.555939pf
C_C4 2 3 0.840164pf
C_C5 3 GND 0.562831pf
C_C6 4 GND 0.555939pf
L_L1 1 5 197.5014nh
R_RL1 5 2 0.568055
L_L2 3 6 197.5014nh
R_RL2 6 4 0.568055
Kn_K1 L_L1 L_L2 0.224933
.ends rw_tag10mm_0

```

Figure 6.8 Subcircuit in PSPICE format.

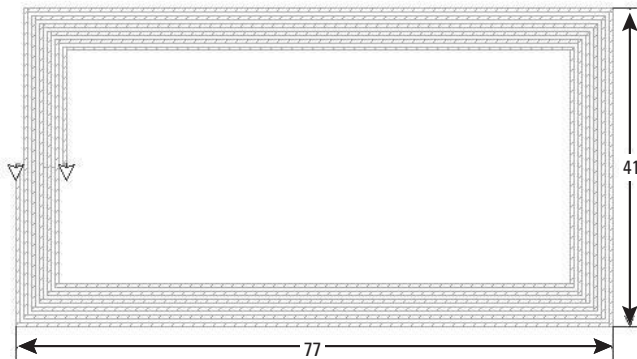


Figure 6.9 A 6-turn coil modeled by Sonnet Lite. File: rw_tag10mm.son.

an enlarged view around the start and end of the winding, including the nearby right angle bend. The line width and spacing are both 0.5 mm. The coil is 77 mm wide and 41 mm long. Vias are at both the start and end (small triangles). An internal port is set (Figure 6.11) in the lower layer (Level 1). And, as before, we change the PI Model Lmax, but this time to 10,000.

We want to put a via on the edge of a cell upward from Level 1 to Level 0. To make sure it goes up (instead of down), select Tools->Add Via->Up One Level. Next, select Tools->Add Via->Edge Via, then click on the edge of the short line drawn in Level 1.

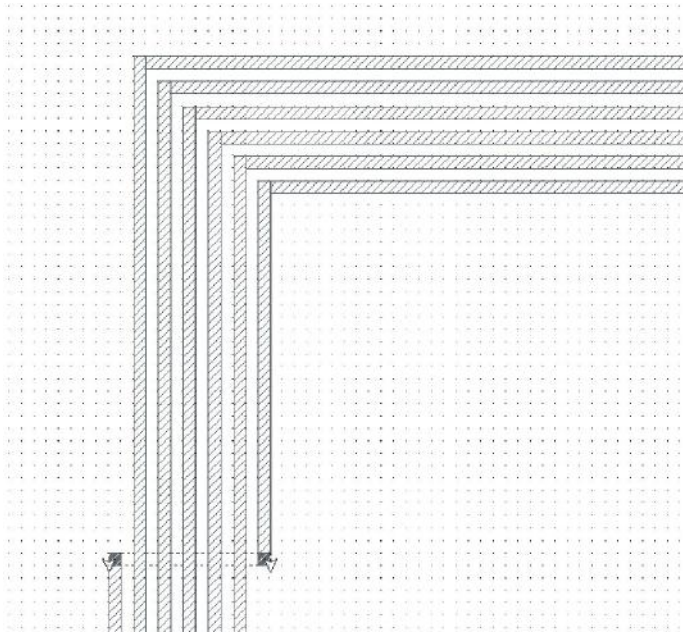


Figure 6.10 Enlarged view around the port.

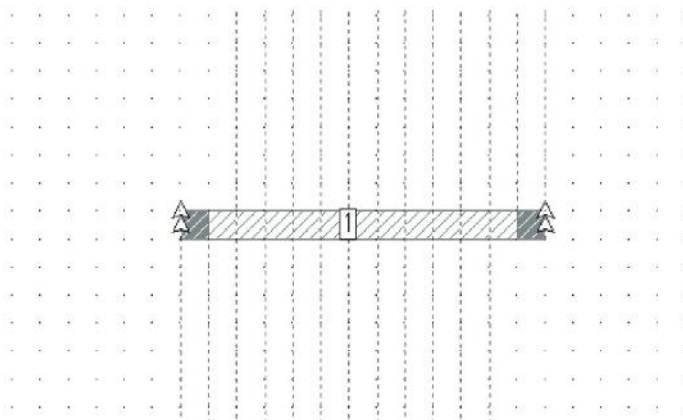


Figure 6.11 Internal port on Level 1.

As for an antenna simulation in Sonnet Lite, it is important to pay attention to the size of the Box. In the case of the one-half wavelength dipole antenna in Chapter 3, we put the antenna about one wavelength away from

all Box sidewalls. As for the top and bottom of the simulation space, we set free space by setting Top Metal and Bottom Metal to Free Space; however the models in Figures 6.5 and 6.9 do not follow these guidelines. One wavelength is way too long.

In addition, the memory usage of Sonnet Lite cannot exceed 16 MB. Fortunately, it is possible to use a small simulation space for coils, because we only need to analyze the magnetic field in and around the coil. The coil is about 1/1,000 of a wavelength in size. For practical purposes, there is no far field. So, we find that setting the Sonnet Lite box size larger than one or two coil diameters allows good results.

We use aluminum of $10\text{ }\mu\text{m}$ thick for the coil conductor (Figure 6.12). This is set by selecting Circuit->Metal Types, by clicking the Add button, and selecting Aluminum from the Global Library in Select metal from library. Figure 6.13 shows the contents of the file obtained by setting SPICE subcircuit output after selecting Analysis->Output Files. The result is an equivalent circuit

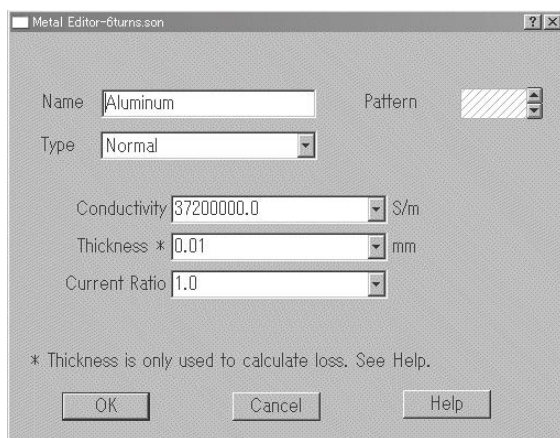


Figure 6.12 Coil metal is set to aluminum $10\text{ }\mu\text{m}$ thick.

```
* Analysis frequencies: 13.3, 14.62 MHz
.subckt 6turns_3 1 GND
C_C1 1 GND 1.153935pf
L_L1 1 2 4466.234nh
R_RL1 2 GND 6.868912
.ends 6turns_3
```

Figure 6.13 Synthesized SPICE subcircuit model.

for the vicinity of 13.56 MHz. We can see that this 6-turn loop at around 13.56 MHz is represented as a circuit that has values of L , C , and R as shown in Figure 6.14.

The value of self-inductance L between node 1 and node 2 shows the L of the 6-turn coil in the tag. C , between node 1 and node 0 (GND) is the slight amount of capacitance caused by the capacitive coupling between lines. And because we are using a real metal, aluminum, with nonzero resistance, the value R also appears. This model is then used to design RFID tags using this coil. The details of such a design depend on the chip selected, and are typically provided by the chip manufacturer.

6.2 UHF RFID Tag Antennas

Different frequencies have different characteristics. Above, we have explored one of the lower frequencies used for RFID. Next, we explore a higher frequency.

6.2.1 Application of an RFID Tag

In experiments involving attaching RFID tags on various items at the supermarket, the advantages and disadvantages of RFID tags at 13.56 and 2.45 GHz became clear.

It was found that 13.56 MHz RFID tags could be read on all items except things like metallic cans used for beer and soda. Meanwhile, 2.45-GHz RFID tags could be read less than one-third of the time as compared to 13.56 MHz tags. When attaching tags directly on items with high water content, such as a full plastic bottle or a Japanese radish, most tags could not be read at all.

The 900-MHz band is an intermediate band. RFID tags for this frequency can be read over a much longer distance than 13.56 MHz and are not affected as much by water as compared to 2.45 GHz.

Industry is developing a system which controls and monitors the entire supply chain, for shipment, delivery, and sale by attaching RFID tags on products during manufacturing.



Figure 6.14 Schematic diagram of the SPICE subcircuit.

6.2.2 Half-Wavelength Dipole Antenna for the UHF Band

A half-wavelength dipole antenna for 950 MHz is about 16 cm long. This might be an entirely acceptable size to place on some product boxes. However, it must be miniaturized if it is to be used on small products.

Figure 6.15 is an RFID tag with a linear dipole antenna; the size is 1 inch (2.54 cm) $\times$ 6 inches (15.24 cm). The element structure in the center to which the IC chip is attached functions as an impedance-matching network.

As we learned in Chapter 4, when increasing the permittivity of a substrate, we can reduce the size of an electric field detection type antenna. However, we typically produce RFID tags by printing antennas on PET films or thin paper and then affixing foils. Once the substrate material is set, we must use other miniaturization methods.

For example, what happens if we bend the dipole as shown in Figure 6.16? Leaving the downturned leg length fixed at 2.6 cm, we parameterize the horizontal length, just as we did for the linear dipole in Chapter 3.

We assume that the dielectric is a thin sheet of PET plastic 1 mm thick with a relative permittivity of 2.2 and $\tan\delta = 0.002$. After running a parameter sweep, we obtain the reflection coefficient shown in Figure 6.17. When the horizontal element is 9 cm long it resonates at about 950 MHz. The total length of the complete dipole is similar to that of the linear dipole in Chapter 3;

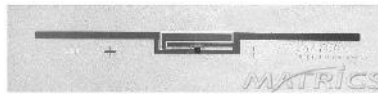


Figure 6.15 Example of an UHF RFID tag, a simple dipole antenna.

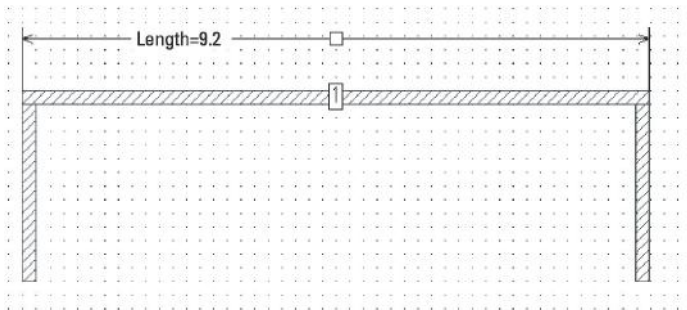


Figure 6.16 Parameterized bent dipole. File: bentdp_param.son.

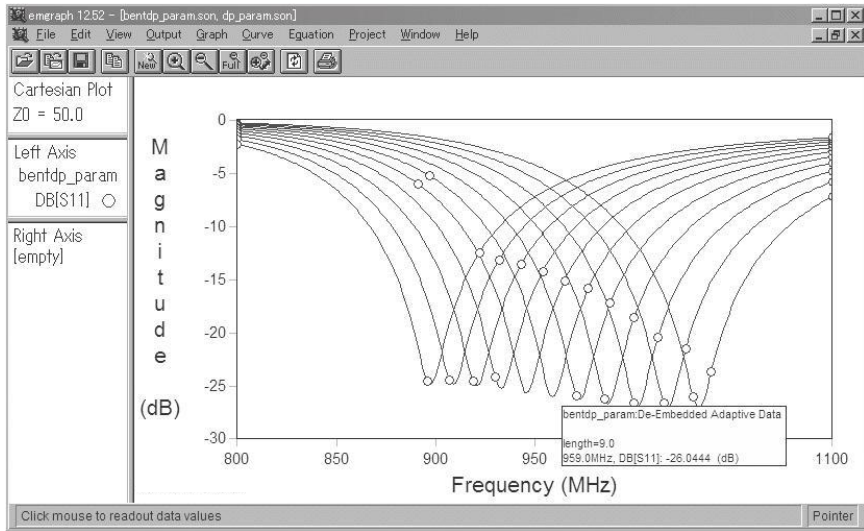


Figure 6.17 Result of the parameter sweep. With the horizontal portion of the dipole at 9 cm long, we can realize operation at 950 MHz.

however, the horizontal dimension needed for mounting is reduced by 60%. As we have seen, we can change the shape of an antenna, and still make it resonate at the desired frequency. There is a trade-off, however. The bandwidth is slightly narrower (Figure 6.18).

6.2.3 Broadband Techniques

Ideally, RFID tags use broadband antennas to support three bands, for Europe (868 MHz), the United States (915 MHz), and Japan (953 MHz). Tags are affixed on various dielectrics such as a carton box, a wooden box, glass, and so on. It would be highly desirable to use one antenna design worldwide, even when affixed to any material. We approach this by designing broadband so that resonant frequency shifts caused by the wavelength shortening effect do not disable the tag.

As for amateur radio, multiple narrow operating frequency bands are available, so multiband operation with one antenna is highly desirable. For example, as shown in Figure 6.19, when you feed different length dipoles all connected to one feed point with one cable, we have a multiband dipole antenna. This type of antenna has been popular for decades.

Figure 6.20 is a dipole antenna with different length elements using this same idea. It was hoped to realize a broadband response because the two dipoles are resonant at two frequencies. Figure 6.21 shows the reflection coefficient. We

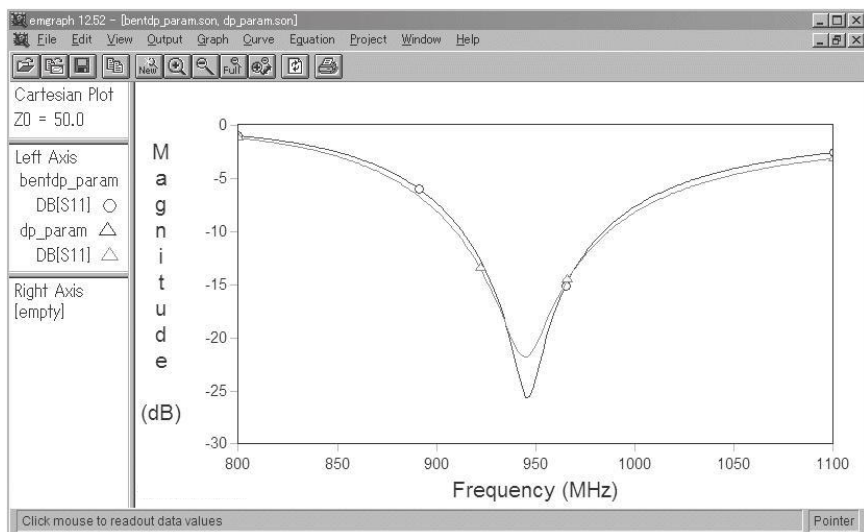


Figure 6.18 Reflection coefficient of the bent dipole (small circle marker) and the simple dipole antenna.

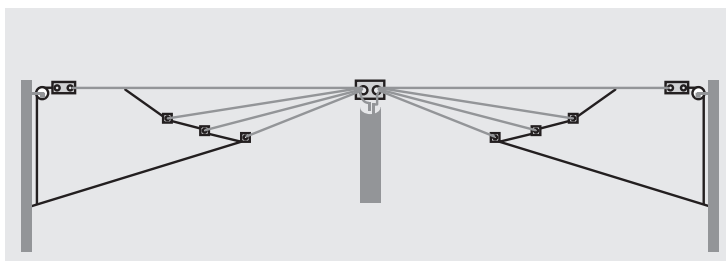


Figure 6.19 Multiband dipole antenna for amateur radio.

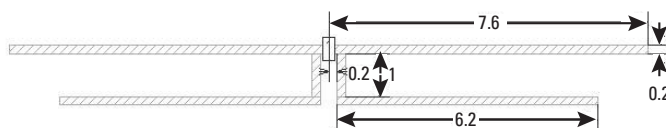


Figure 6.20 Dual-band dipole antenna with different length elements. File: wide_dp.son.

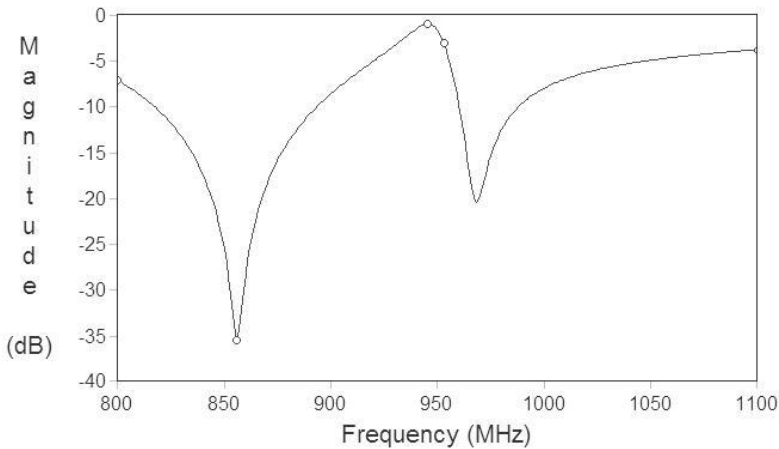


Figure 6.21 Reflection coefficient of the dual band dipole. Large reflection is seen at around 945 MHz.

see that this antenna does indeed have low reflection at two frequencies, 856 and 968 MHz as expected. Unfortunately, in between, at around 945 MHz, we see large reflection, and it is not so good at 915 MHz either. When the tag's resonant frequency is shifted by the surrounding environment, the entire response moves around, very possibly putting high reflection right at the frequency we might need for operation.

What is happening at 945 MHz to generate this large reflection? To find out, we examine the surface current of this antenna at that frequency (Figure 6.22). We see that the current through the feed point is low (look very close to the small port 1 box), but the current on both wires at the center near the feed point is high. The reason is we have an entire half-wavelength U-shaped resonant dipole on one side of the feed point, and a second one on the other side. Both sides are completely happy all by themselves. Neither side needs to send any current at all through the feed point, making the feed point a low current (and high voltage), high impedance, and high reflection point. Figure 6.23 is a plot of input impedance and it has a peak in the real part R at around 945 MHz where it is over 450Ω .

It is nice that at around 856 MHz and 968 MHz, where the reflection coefficient is low, the real part, R , is near 50Ω with imaginary part X at zero, indicating resonance. At around 945 MHz, X is almost zero as well, indicating a resonance. But R is very high, which indicates an antiresonance (parallel resonance) mode. This would also make an effective antenna, but it would need a high impedance feed. A feed designed for the other two, low impedance, reso-

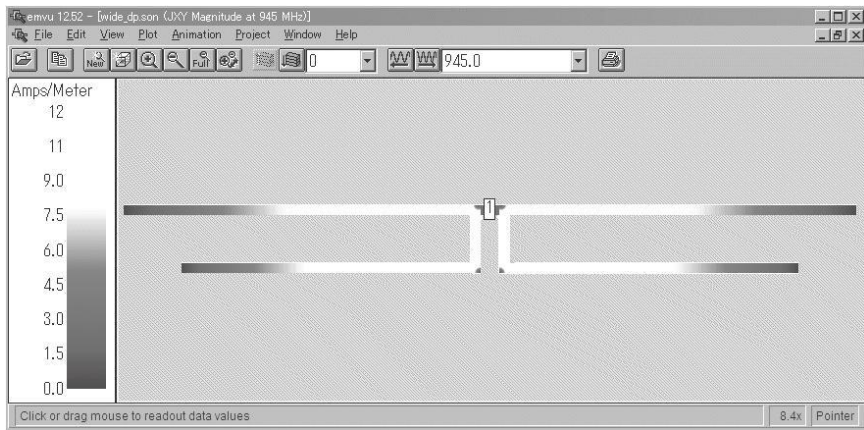


Figure 6.22 Surface current distribution at 945 MHz. Current passing into port is low.

nances does not work. Thus, this antenna cannot realize a broadband response when being fed directly by low impedance cable.

6.2.4 Changing the Element Location

Total input reactance, X , is zero at the antenna resonant frequency. This is the same as in LC resonant circuits. With excitation by a sine wave, the electrical energy accumulating in C and the magnetic energy accumulating in L alternate by 90° phase, and we have the state of resonance.

In a dipole antenna, the electric charges of two poles (as in di-pole) at both ends form C and the magnetic field generated around the current along the wire forms L . Now we review Figure 6.23, we can see that the frequencies where X is zero at 856, 945, and 968 MHz, and that resonance in the middle is an antiresonance (parallel LC) mode between the first and last resonances, which are series LC resonances.

The central antiresonance is caused by one-half wavelength of wire on each side being resonant, with absolutely no need to send current to the other side. We will try to eliminate this problem by rearranging the elements so that these antiresonances are forced to send current through the feed point.

Figure 6.24 is a model whose short element that was at the lower left in Figure 6.22 has been flipped up and over the main dipole into the upper left position. To do this, click on the L-shaped short element, select Modify->Flip, and click the up-down arrow button.

If we were to analyze this dipole as it is, we would see that it is slightly improved but that parallel resonance mode still remains. This makes sense, because the half-wavelength of wire on either side of the feed point still has no

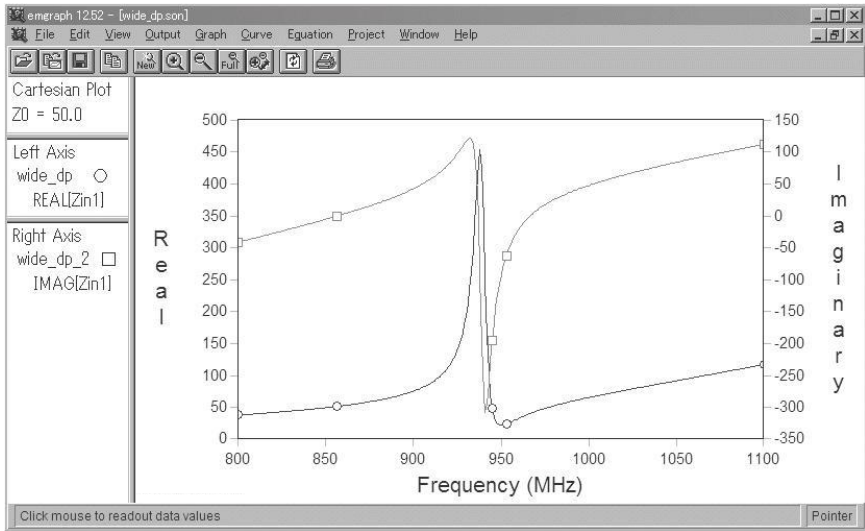


Figure 6.23 Input impedance. R is over 450Ω at 945 MHz.

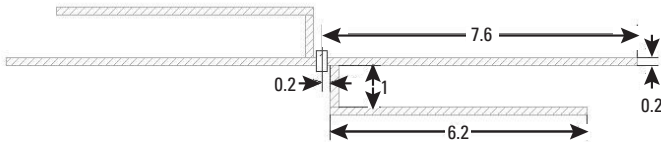


Figure 6.24 The left short dipole element is now flipped up over the main dipole. File: wide_dp2.son.

need for its resonant current to flow through the feed point to the half-wave resonant length of wire on the other side of the feed point.

To force the half-wave resonant current, say, on the left, to flow through the feed line terminals, we just move the short L-shaped elements over to the other side of the feed point (Figure 6.25). Figure 6.26 is the surface current distribution at 919 MHz. We see that the resonant currents on the short elements pass through the feed line port terminals.

Figure 6.27 shows the reflection coefficient of this antenna. We are not finished yet. There is no antiresonance mode so it does seem to be broadband, but the best return loss is 3.5 dB. The input impedance must have been substantially modified. If we check the input impedance, R is indeed 10Ω at most.

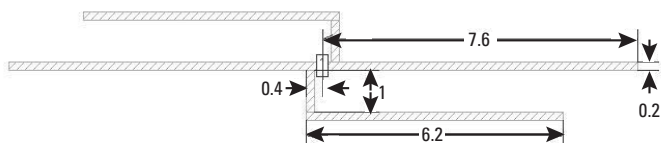


Figure 6.25 Moving the short element's attachment point to the other side of the feed line terminals forces the antiresonant current to flow through the feed terminals.

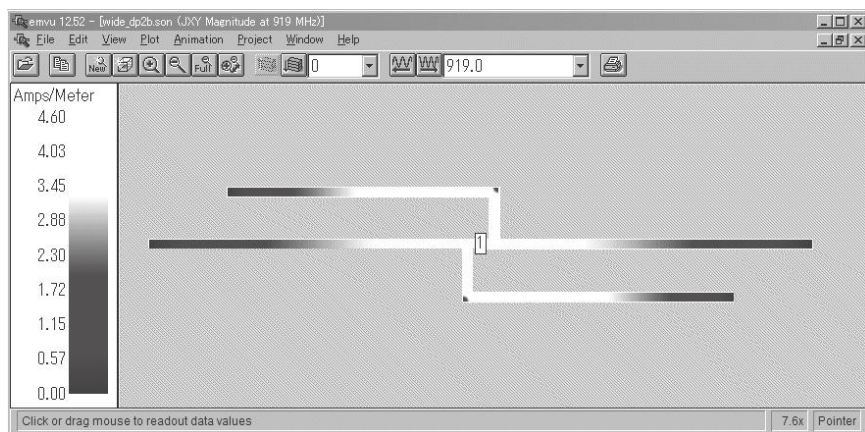


Figure 6.26 Surface current distribution at 919 MHz.

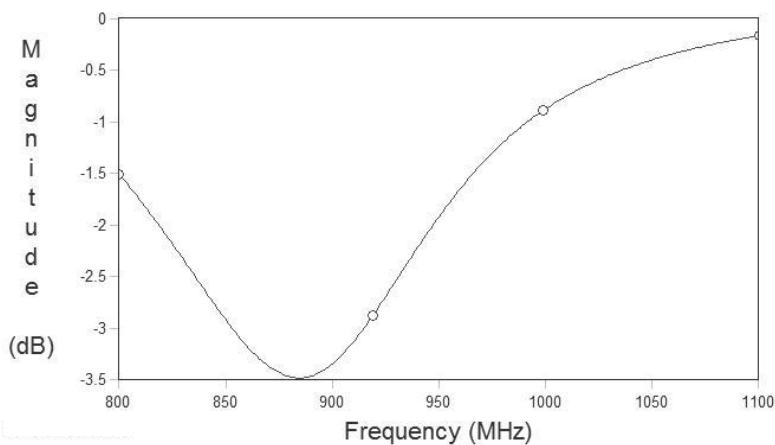


Figure 6.27 Reflection coefficient of the modified model.

Let's not give up yet. Perhaps there is something we can do to match the dipole to 50Ω . Perhaps we should try many dipoles with this same general layout, but many slightly different lengths and separations. Figure 6.28 is a model with the length of the y directed portion of the short element set as a parameter.

The setup procedure is same as we learned in Chapter 3. Select Tools->Add Dimension Parameter->Add Symmetric, click the left button of mouse on the upper right corner of the upper L-shaped element to display a small white square. This sets the first reference point. Then:

1. Click and drag the mouse to select all the points in the upper part of the L-shaped element. This selects all the points that will be moved when the reference point is moved. All of the points are now indicated by small black squares.
2. Press the Enter key. This means you are finished selecting the upper group of points.
3. Click the left button of mouse on the lower left corner of lower L-shaped element, and again, a small white square appears. This is your second reference point.
4. As we did with the upper L element, click and drag to select all the points of the lower part of the element. All the selected points are indicated by small black squares. All of these points will be moved when the second reference point is moved.
5. Press the Enter key.
6. A dialog box is displayed. Type in a convenient variable name.

Set up the simulation just like we did in Chapter 3. Under Analysis ->Setup, add a Parameter Sweep. Select an ABS Sweep from 800 to 1,100 MHz. Then check the check box for varying the parameter and specify the parameter to sweep from 2 to 5 cm in steps of 0.2 cm. Select Project->Analyze to run the analysis.

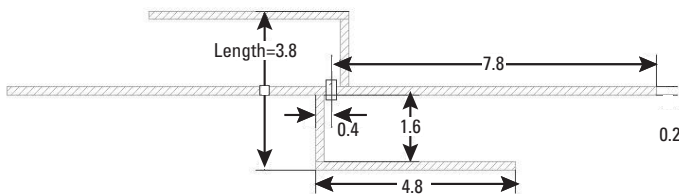


Figure 6.28 Model with the length of the y directed portion of the short element set as a parameter. File: wide_dp3.son.

When the simulation finishes, display a plot of the reflection coefficient. Then, double-click on S11[DB] in the upper left frame and check the Graph All Iterations check box in the lower left and click OK. All reflection coefficients are displayed for all calculated values of our parameter as shown in Figure 6.29.

If you see a response that you want, just move your cursor to point at the desired curve. The value of the reflection coefficient, the frequency, and the value of the parameter are displayed. For example, the curve for length 3.8 cm shows a usable band from 850 to 1,000 MHz with return loss better than 10 dB. This antenna will work in all UHF RFID bands worldwide and even has lots of room for the tag bandwidth to be moved down by nearby dielectric objects. Further optimization of the x directed dipole lengths yields even better performance. This antenna was first described in, “Novel Planar Wideband Omni-directional Quasi Log Periodic Antenna,” presented at the 2005 Asia Pacific Microwave Conference.

6.3 Polarization of Reader and Tag

It would be nice if a tag could be read no matter in what orientation it is held. Polarization is characteristic of electromagnetic waves and refers to the waves, and their antennas, being sensitive to orientation. Do we have a problem here?

6.3.1 UHF RFID Tags

UHF RFID tags are based on the dipole antenna, and communicate by using the far field electromagnetic waves over distances up to about 10m. This is far

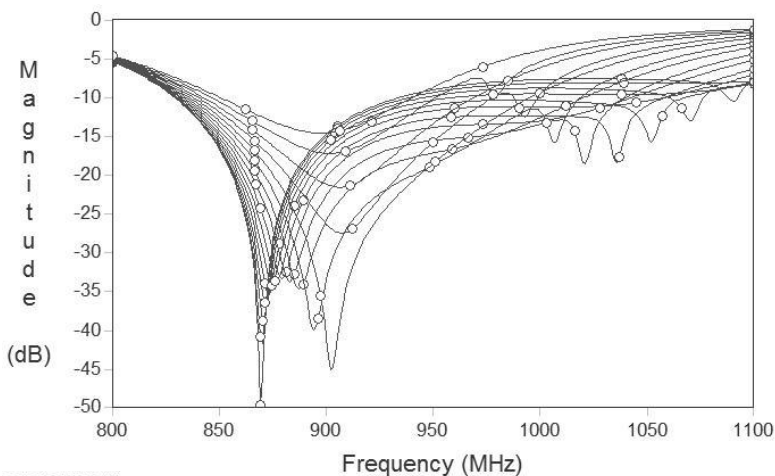


Figure 6.29 Reflection coefficient is displayed for all the values of our parameter.

field because one wavelength at UHF is about 33 mm. So we can imagine attaching tags on the cartons moving along on conveyor belts being read from a long distance. Or we could attach them to pallets of goods, to trucks and fork lifts, and read all the tags as they pass gates, and so forth. It is possible to track goods in the supply chain by attaching tags (in many possible places and orientations) on a carton (Figure 6.30) and in parts management and product management by attaching them on assembled parts at factories.

A horizontal dipole antenna can receive a signal transmitted by another horizontal dipole. The signal transmitted by the horizontal dipole is called horizontally polarized. If we turn one of the dipoles by 90° , then it is vertically polarized, and little if any signal is received.

When we work out the mathematics, it turns out that a horizontally polarized signal has a horizontal electric field. Of course, a vertical dipole transmits a vertically polarized signal. For good reception, the transmitting and receiving antenna should be the same, or almost the same, polarization.

It is also possible to combine a horizontally plus a vertically polarized wave so that the electric field vector spins around at the same frequency as the electromagnetic wave. This is called a circularly polarized wave. In fact, depending on how the two waves are combined, the electric field vector can spin clockwise or counterclockwise. An important note: what clockwise and counterclockwise mean depends on whether you are viewing the wave from the transmitter or from the receiver. Both points of view are used in practice. The terms “right-hand circular” and “left-hand circular” are also used. A patch antenna can be easily set up to transmit and receive circularly polarized waves.

In Figure 6.30, when a horizontally polarized wave comes from front left, tag (1) receives best. Also (3) and (7) can receive, but (2) and (4), which are vertically polarized, can hardly receive anything. Note that we are simplifying a

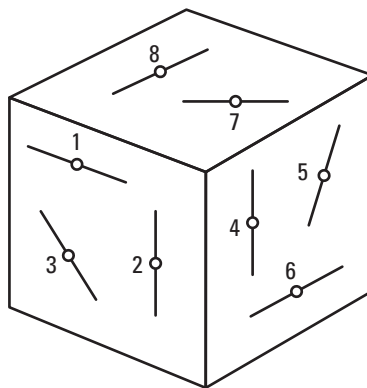


Figure 6.30 RFID tags scattered over a cardboard box.

little bit here, assuming that the box and the tags do not influence the polarization of the wave being received.

If the reader is circularly polarized, then the transmitted wave is half vertical and half horizontal. A vertical antenna receives the vertical half. A horizontal antenna receives the horizontal half. In this case, only tags (6) and (8) no longer receive the circularly polarized wave. This is because the ends of their dipole are pointed right at the reader. Again, in practice, the box interacts with and modifies the transmitted wave. We can imagine, for example, that tag (5), on the back of the box, receives no signal of any kind, and thus cannot be read.

Figure 6.31 shows an example of an antenna formed from two dipoles in the form of a Buddhist cross manufactured by Matrics™ (presently Motorola Symbol). The elements are folded to minimize area. The current near the IC is strong and is available for both vertically and horizontally polarized waves.

With orthogonal dipole elements, we have a crossed dipole. The tag in Figure 6.31 can receive both horizontal and vertical waves coming from a reader in any direction, even when attached on a carton as shown in Figure 6.32. Well, at least as long as some signal, of any polarization, can get to the tag.

6.3.2 Buddhist Cross-Shaped RFID Tag

Figure 6.33 shows an antenna model in the approximate shape of a Buddhist cross. It is a rough drawing, and analysis shows that it can cover the worldwide UHF RFID bands.

RFID tags for UHF are attached on a carton box and their resonant frequencies shift to lower frequencies under the influence of nearby dielectric objects. Thus, in general, we should design RFID tag antennas to resonate at slightly higher frequencies when they are in free space.

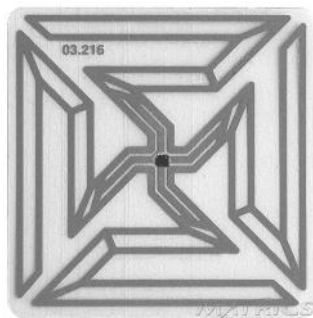


Figure 6.31 Modified crossed dipole antenna, from Matrics.

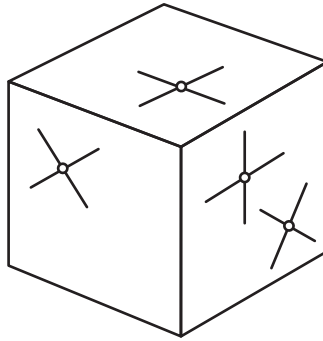


Figure 6.32 Crossed dipole tags on a cardboard box.

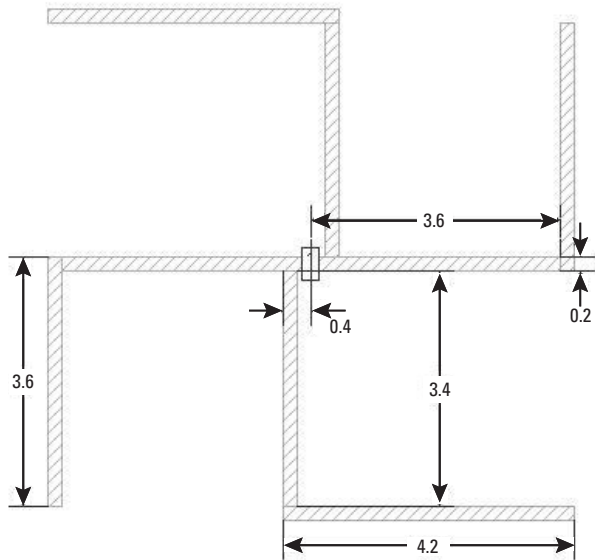


Figure 6.33 Buddhist cross dipole RFID tag, file: cross_dp.son.

6.4 Radiation of Circular Polarization from Patch Antenna

It looks like circular polarization can be helpful. Let's see if we can get circular polarization from a patch antenna.

6.4.1 Simulation Model of a Circularly Polarized Patch Antenna

Figure 6.34 shows a feeding method for a patch antenna that can generate a circularly polarized wave. The left feed line is longer by a one-quarter wavelength and the patch is fed by two microstrip lines (MSLs) connecting on adjacent sides of the patch. One patch plays a dual role as two antenna elements, and antenna elements are excited simultaneously, but they are orthogonal to each other, so both a horizontally and a vertically polarized wave is generated. When the delay between the two waves is 90° (remember the one-quarter wavelength longer feed line on the left), the sum of the two waves form a circularly polarized wave.

Figure 6.35 is an example of a circularly polarized wave patch antenna that is included in Sonnet Lite. Select Help->Examples, and you can find the RHP example in the Antennas section in the Application Examples. You can load the example into Sonnet Lite directly from the help page, or from file cross_dp.con on the disc.

6.4.2 Right-Handed and Left-Handed Polarization

In actual wireless communications, electromagnetic waves pass through various dielectrics such as buildings and ground to be reflected, refracted, diffracted, and so forth, and arrive as a mixture of different polarizations from different directions, and even with different delays. So in the real, physical world, we rarely see a simple, clean, monopolarized wave.

A circularly polarized wave can be obtained by time-shifting a vertically polarized wave and a horizontally polarized wave by 90° with respect to each other, and letting them propagate simultaneously as shown in Figure 6.36. The electric field is composed of both field vectors, as shown by the arrows drawn

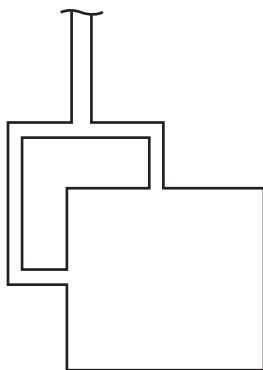


Figure 6.34 Circularly polarized patch antenna.

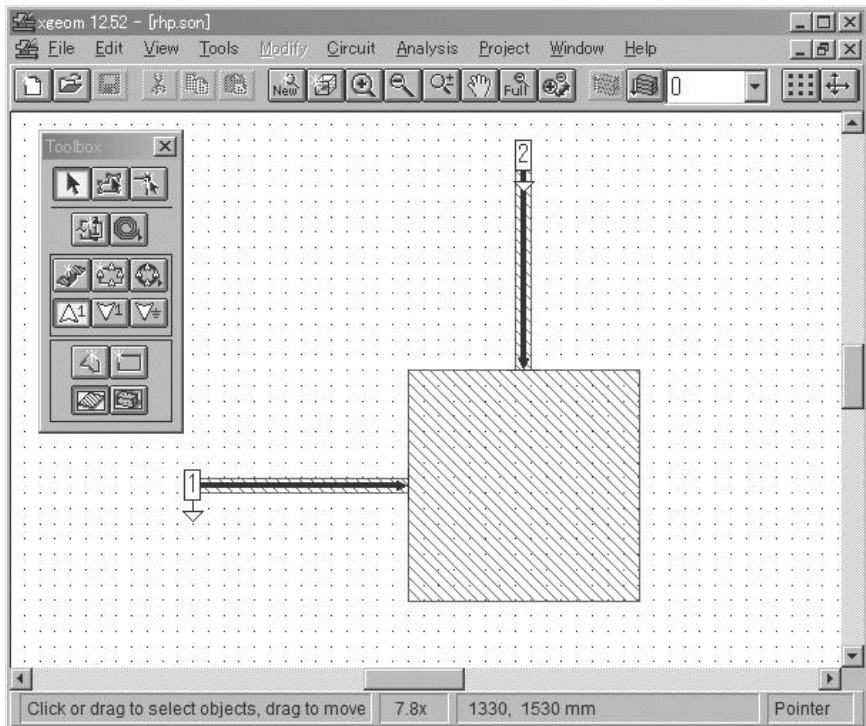


Figure 6.35 Sonnet example model of a circularly polarized patch antenna.

on a transverse plane surface in Figure 6.36, which shows that the electric field vector is circling around as we go through one cycle of the wave.

In Figure 6.36, only the electric field is drawn, and the electromagnetic wave is traveling in $+z$ direction. As viewed from the transmitter (indicated by the eye), at time instant 1, the electric field vector is horizontal, pointed along the x axis at the transverse plane indicated in the figure. One-quarter cycle later, at time instant 2, the wave has traveled one-quarter wavelength further along the z axis to give us an electric field vector that is vertical, along the y axis. And so it goes. From the viewpoint of the transmitter, the electric field vector is rotating clockwise, and we call it right-handed polarization.

6.5 Prediction of Communication Distance

It would be nice to at least take an educated guess as to how far a transmitter/receiver pair can function before we build it. There is indeed a lot of randomness that cannot be controlled, but we can do better than just hope it will work.

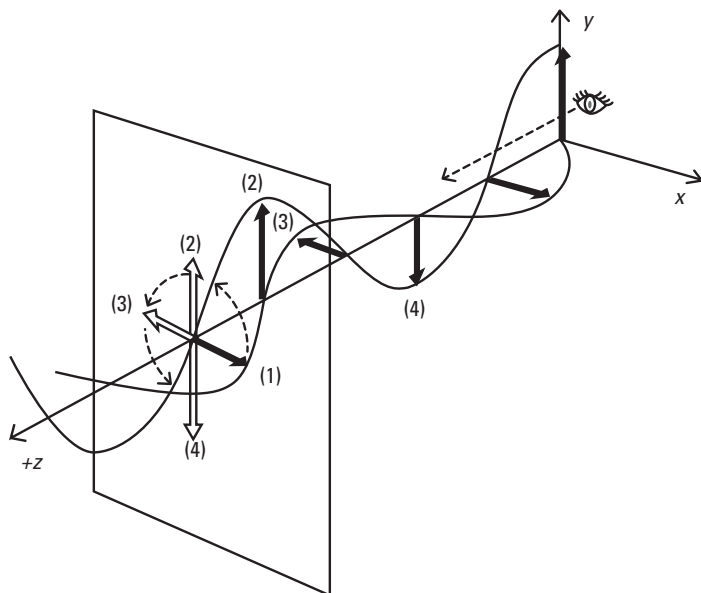


Figure 6.36 A circularly polarized wave can be obtained by adding two orthogonal linearly polarized waves with phase difference of 90° .

6.5.1 Communication Distance of UHF RFID Tags

Figure 6.37 shows a transmitting antenna and a receiving antenna in free space separated from each other by distance r . The power P_t is applied to the transmitting antenna. The antenna gain (discussed in Chapter 7) is G_t . The power density P_d of the wave arriving at the receiving antenna assuming a spherical wave is then

$$P_d = \frac{P_t G_t}{4\pi r^2}$$

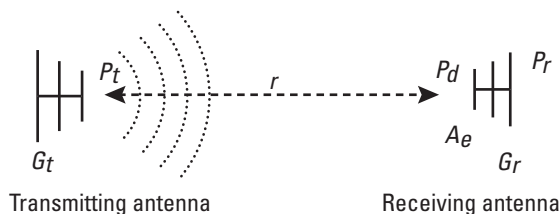


Figure 6.37 A transmitting antenna and a receiving antenna in free space.

For those who are comfortable with mathematics, you will realize that the above equation just says that the power from the transmitting antenna is simply spread out on the surface of a sphere.

When the minimum power receivable by a receiver (sometimes called minimum discernable signal or minimum detectable signal) is P_r and the gain of a receiving antenna is G_r , the expression for the maximum distance r is

$$r = \lambda \sqrt{\frac{P_t G_t G_r}{(4\pi)^2 P_r}}$$

The expression to approximate the communication distance between a UHF RFID tag and its reader, stated in words, is:

$$\text{Distance [m]} = \lambda \sqrt{\frac{\text{Transmission EIRP [W]} \times \text{Gain of tag} \times \text{Polarized waveloss}}{(4\pi)^2 \times \text{Minimum power of tag [W]}}}$$

EIRP stands for *effective isotropic radiated power*. An isotropic antenna is an antenna that radiates equally well in all directions. It does not, in fact it cannot, physically exist. It is useful however, as a theoretical standard to rate other antennas. So, EIRP means the amount of power you would have to pump into this fictitious antenna to get the same power density at our receiver as we get with the real antenna. Thus, if our real antenna has a gain of 3 dBi (dB over isotropic), then 1W into our actual antenna gives the same signal as 2W into an isotropic antenna, and we have 2W EIRP.

Polarization loss is 1.0 if there is no loss, and it is 0.0 if there is complete and total loss of signal due to polarization mismatch. If you know the difference in the angle between the transmitted polarization angle, and the polarization angle that the receiver expects, just square the cosine of that angle to get the polarization loss. For dipoles in free space, the mismatch angle is just the angle between the dipoles as viewed along a line connecting the dipoles.

If there are some metal objects around, RFID systems can sometimes communicate over longer distances than expected because of reflection, refraction, and so forth. Unfortunately, it can also be a lot shorter. Thus, as we mentioned above, there is a lot of randomness. This is where the field of statistics can be applied. The mathematics are far too involved to go into here, but for those who are so inclined, this can be an extremely rewarding application of an advanced field of mathematics.

7

Determination of Antenna Characteristics by Using EM Simulators

7.1 Radiation Efficiency of Antennas

Nothing is perfect, including antennas. To do engineering, we must put a number on whatever we are interested in, and then try to design our systems so that that number is better. Here we consider a number that is especially critical for small antennas; the number we call radiation efficiency.

7.1.1 Definition of Radiation Efficiency

Radiation efficiency, η (eta), refers to the percentage of power actually radiated by an antenna. It is a ratio, usually expressed in percent, defined by the following equations:

$$\eta = \frac{P_{rad}}{P_{in}} = \frac{R_{rad}}{R_{in}} = \frac{R_{rad}}{(R_{rad} + R_{loss})} \times 100[\%]$$

Here P_{rad} is radiated power, P_{in} is input power, R_{rad} is radiation resistance, R_{in} is input resistance, and R_{loss} is loss resistance.

The unit of radiation resistance (R_{rad}) is the ohm; however it does not refer to the ohmic loss caused by the metal of an antenna. Radiation resistance R_{rad} is defined by the following equation:

$$R_{rad} = \frac{P_{rad}}{|I|^2}$$

Here, I is the current at the feed point. In an electromagnetic simulation, the radiation resistance R_{rad} corresponds to the real part of the input impedance, R , of an antenna built using perfectly lossless materials. The input resistance of such a lossless antenna is due entirely to power loss due to radiation.

Then, when we build the antenna from real, lossy, resistive materials, the input resistance R is now the sum of the radiation resistance R_{rad} (lossless) and loss resistance of the antenna R_{loss} .

This loss resistance comes from conductor resistance, loss in any ground or earth near the antenna, dielectric loss, and so on. Using better materials is only one (usually expensive) way of improving the radiation efficiency of an antenna. Notice the above equation for efficiency. We can also work to increase the radiation resistance, too. Well, anyway, as long as whatever we do does not also increase the loss resistance.

The R_{rad} of a one-half wavelength dipole in free space is 73Ω . This is typically much larger than the loss resistance, so we can realize very high-efficiency antennas with ordinary metal wire.

Figure 7.1 shows an example of a simulation result for a dipole antenna that includes evaluation of the radiation efficiency. Keep in mind, not all commercial simulators evaluate radiation efficiency. If this is the case, you can calculate η by the following equation that includes the actual antenna gain, G_a , and the directive gain G_d , both in dB. We give an example and explain more about this equation in Section 7.1.4.

$$\eta[\%] = 100 \times 10^{(G_a - G_d)/10}$$

7.1.2 Measuring Radiation Efficiency

As the input power is easily measured, we could calculate efficiency most easily using the first equation above, if only we could measure the total radiated power. Unfortunately, it is virtually impossible to gather and measure all the power radiated out into space.

Consequently, the Wheeler cap was invented (<http://www.rfcafe.com/references/articles/Antenna-Efficiency-Wheeler-Cap/Efficiency-Measurement-Antenna-Wheeler-Cap.htm>). Figure 7.2 shows a Wheeler cap developed at Universitat Politècnica de Catalunya (UPC) and Ecole Polytechnique Fédérale de Lausanne (EPFL) by José M. González, Jordi Romeu, Eugenia Cabot, and Juan R. Mosig. It is a hollow metallic sphere. When an antenna is small enough to be placed inside, radiated power can be measured over a wide range.

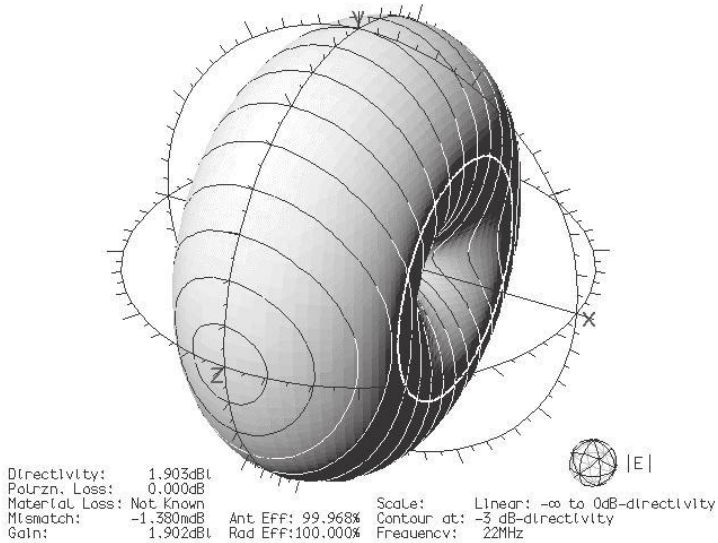


Figure 7.1 Simulation result for a dipole antenna including calculation of the radiation efficiency.



Figure 7.2 The Wheeler cap allows measurement of total radiated power. (From: "Task 4.3 Final Report" available at <http://www.tsc.upc.edu/fractalcoms/>).

The Wheeler method, using the Wheeler cap, is one technique to measure radiation efficiency. With the antenna inside the Wheeler cap, the real part of the input impedance is just R_{loss} . The metal cap prohibits radiation, so the only loss is loss due to the antenna materials.

To complete the radiation efficiency measurement, we measure the real part of the antenna input impedance in free space. Now, the real part is the radiation resistance plus the loss resistance. In practice, the reflection coefficient of the antenna input is measured. The radiation efficiency η is then

$$\eta = 1 - \frac{1 - |\Gamma_w|^2}{1 - |\Gamma_f|^2}$$

where Γ_w is the antenna reflection coefficient measured with the antenna inside the Wheeler cap and Γ_f is that measured in free space. In this method, electromagnetic waves are radiated inside a sealed metal sphere, so this method cannot be used at any resonant frequencies of the cavity. If the shielding is imperfect in anyway, say, if it radiates electromagnetic waves from a slit, then the value of R_{loss} is inaccurate.

In addition, if the radiation efficiency of our antenna exceeds 80%, the value of $R_{rad} + R_{loss}$ becomes nearly equal to R_{rad} all by itself and measurement error can become large. Since it is almost impossible to realize over 80% radiation efficiency for miniaturized antennas, the Wheeler method seems well suited to their measurement.

7.1.3 A Method for Calculating Efficiency Using EM Simulators

Measuring the Q (quality) of a resonator is a well-known method of measuring the loss. An antenna is a resonator, too. In fact, unlike the low loss, narrowband resonators in a typical filter, antenna Q is usually fairly low. This actually makes it fairly easy to measure antenna loss. When we use EM analysis, we can easily compare the Q of an antenna built with lossless materials to that of the same antenna built with real, lossy materials. We can figure out the antenna efficiency by comparing the two results. Of course, we cannot build an antenna with ideal lossless materials, so this method does not work for actual measurements. However, it works just fine when we use EM simulators.

In Figure 7.3, we see a bent dipole antenna modeled in Sonnet Lite. First, by simulating with all lossless metals, we get the reflection coefficient, S_{11} (Figure 7.4).

The Q is obtained by first finding the frequency of the minimum reflection coefficient. At this frequency, the reflection coefficient Γ_0 is 0.045. Complete reflection is a reflection coefficient of 1.0. The average of these two values is 0.52. The bandwidth is determined at this frequency, as indicated in Figure 7.4. This bandwidth represents the sharpness of the resonance. It is related closely to the Q of the resonant antenna. It is

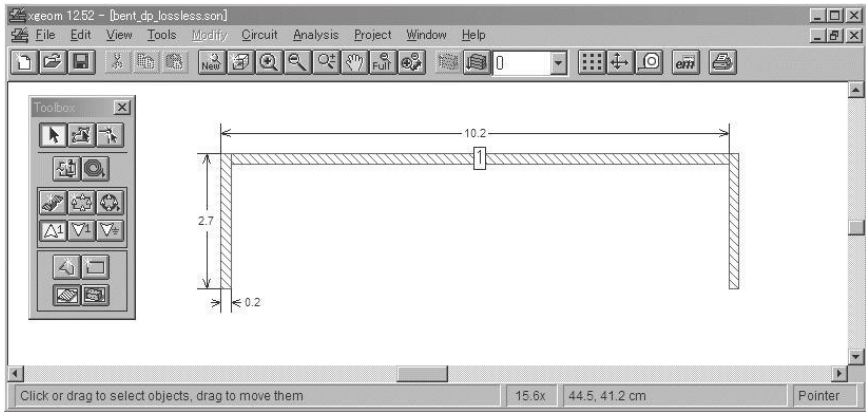


Figure 7.3 Model of a bent dipole antenna. File: bent_dp_lossless.son.

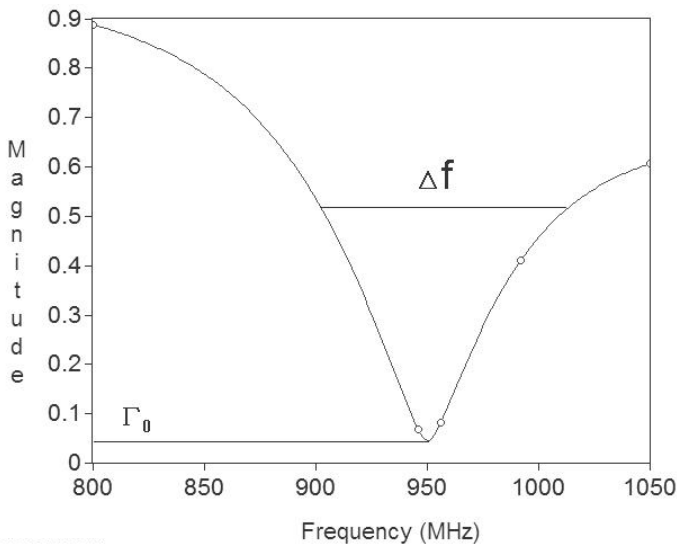


Figure 7.4 Reflection coefficient of S_{11} for a bent dipole antenna with all lossless metals.

$$\frac{\Delta f}{f_0} = \frac{1}{Q}$$

where Δf is the half-power bandwidth, and f_0 is the center frequency. The radiation efficiency is determined by the ratio of Q with loss to the Q of the lossless

antenna. Referring to the equation above, it is also the ratio of the bandwidth without loss to the bandwidth with loss.

Figure 7.5 shows the calculated reflection coefficients for both an aluminum dipole and a lossless dipole. Figure 7.6 shows the graph magnified.

Reading out the bandwidths, we have 112.05 and 112.38 MHz. Their ratio is the radiation efficiency of the dipole antenna in free space, which is pretty good.

$$\eta = \frac{112.05}{112.38} \times 100 = 99.7\%$$

7.1.4 Radiation Efficiency of Patch Antennas

Figure 7.7 is the patch antenna model fed by a via port, which we simulated in Chapter 4. The dielectric is FR4 with a relative permittivity of 4.9 and loss tangent of 0.025, and 1 mm thick. As patch antennas radiate by means of the strong electric fields from two edges, we might worry that with only 1-mm thickness, the radiation efficiency might not be so good. Let's check this possibility quantitatively.

We discuss antenna gain more in the next section, but a quick introduction will help in understanding what we are doing next. When the transmitting antenna is pointed directly at the receiver, the actual gain determines how

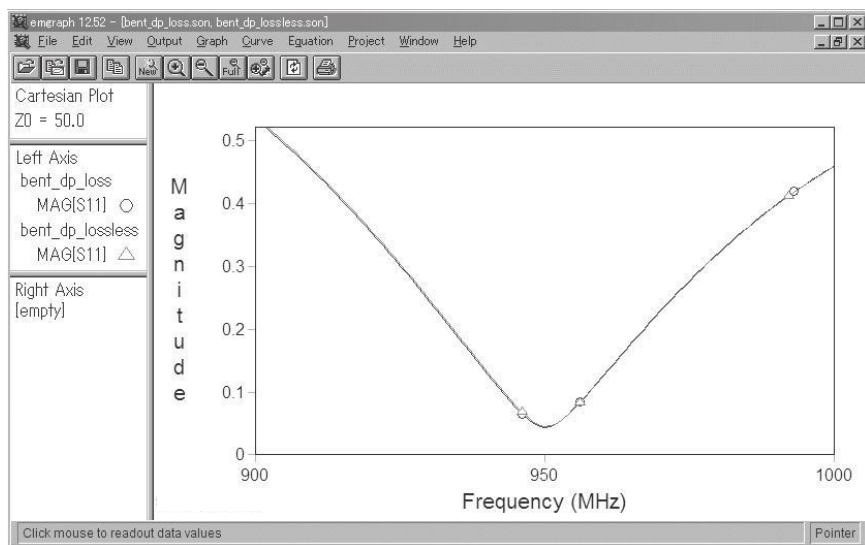


Figure 7.5 Reflection coefficients for an aluminum dipole and a lossless dipole.

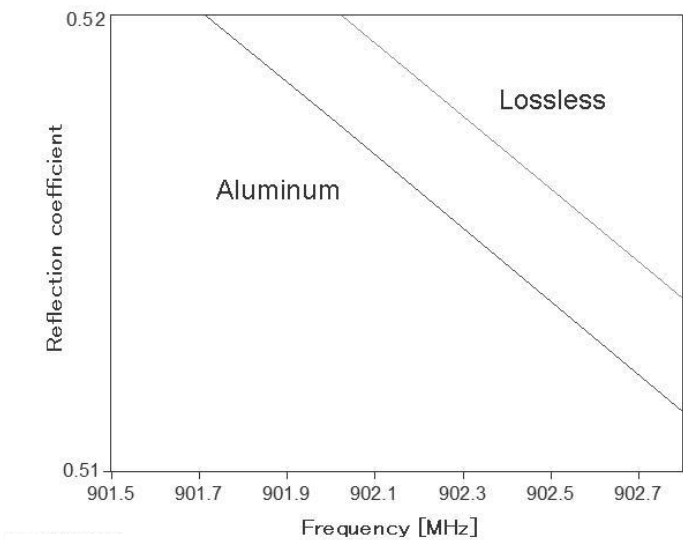


Figure 7.6 Magnified portion of the graph shown in Figure 7.5.

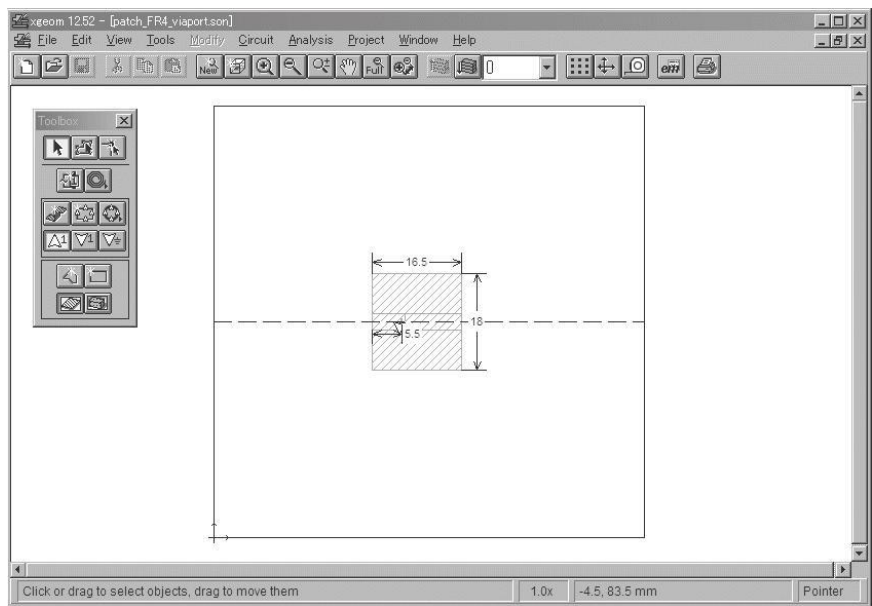


Figure 7.7 Patch antenna model fed by a via port. File: patch_FR4_viaport.son.

strong the transmitted signal will be at the receiver. There is no need to know anything else about the antenna pattern.

In contrast, the directive gain requires knowledge of the entire antenna pattern. It gives an indication about how directive the antenna is. Remember the Beverage antenna in Section 5.1.3? It is a highly directive, narrowbeam, antenna. It can strongly reject signals off the side or off the back. But because it uses the earth for a ground return, it is a very lossy antenna. Thus, the actual gain is very low, but the directive gain is very high. This means that the Beverage has a very low efficiency. If an antenna is lossless, the actual gain equals the directive gain.

In this example, we determine the radiation efficiency of our patch antenna by determining the actual and the directive gain. This requires a detailed evaluation of the far field. Note, however, that the far-field module is not available in Sonnet Lite. Figure 7.8 is the far-field radiation pattern of this patch antenna.

In Figure 7.8, the plot is set up to display the actual gain, see the upper left-hand corner. We see $G_a = 1.7$ dB. By selecting Graph->Normalization...->Directive Gain(dB), then we can read out the Directive Gain. The directive gain is $G_d = 6.9$ dBi. Now we calculate the radiation efficiency as follows.

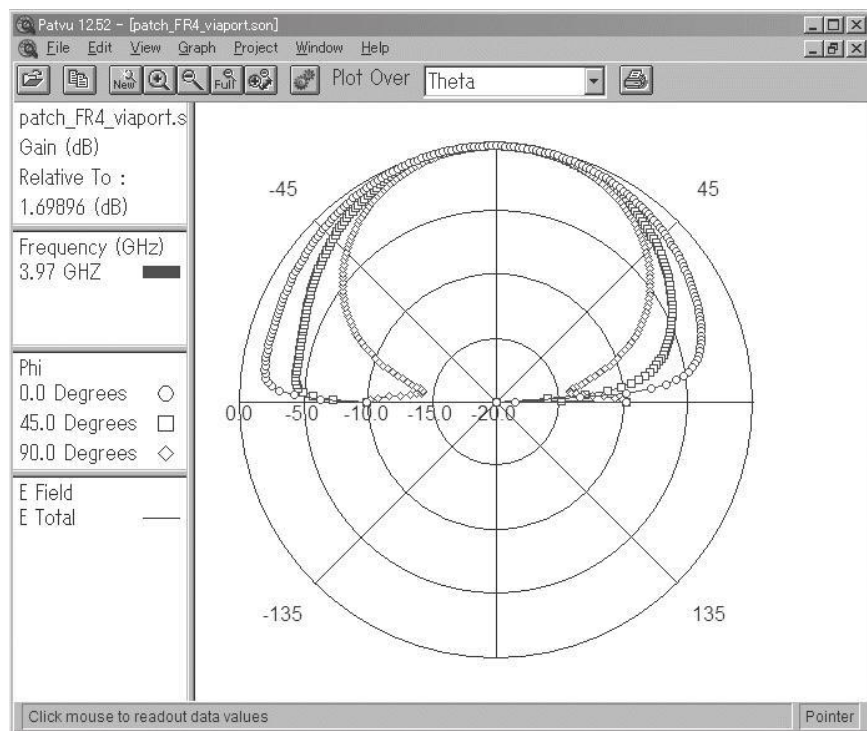


Figure 7.8 Far-field radiation pattern of this patch antenna.

$$\eta[\%] = 100 \times 10^{(G_d - G_a)/10} = 100 \times 10^{(1.7 - 6.9)/10} = 30\%$$

This value means that if we feed 1W into the antenna, it radiates only 0.3W. This is because of resistive loss from current flowing through the metal of the patch, and dielectric loss as the high electric field at the patch edges shake the electron clouds around atoms of the dielectric back and forth. To improve the radiation efficiency, standard practice is to increase the thickness of the substrate. Let's double it to 2 mm and see what happens (Figure 7.9).

With a dielectric substrate now 2 mm thick, the length of this patch, shown in Figure 7.7, should now be made 1 mm shorter (Figure 7.9), so as to keep about the same resonant frequency. Figure 7.10 shows the far field radiation pattern of this patch antenna. The efficiency calculated from the actual and the directive gain is now 46%. Making the substrate thicker did indeed increase the antenna efficiency.

7.2 Antenna Gain

We have already started using antenna gain to determine antenna efficiency. Gain is an important concept. Let's make sure we have a solid understanding.

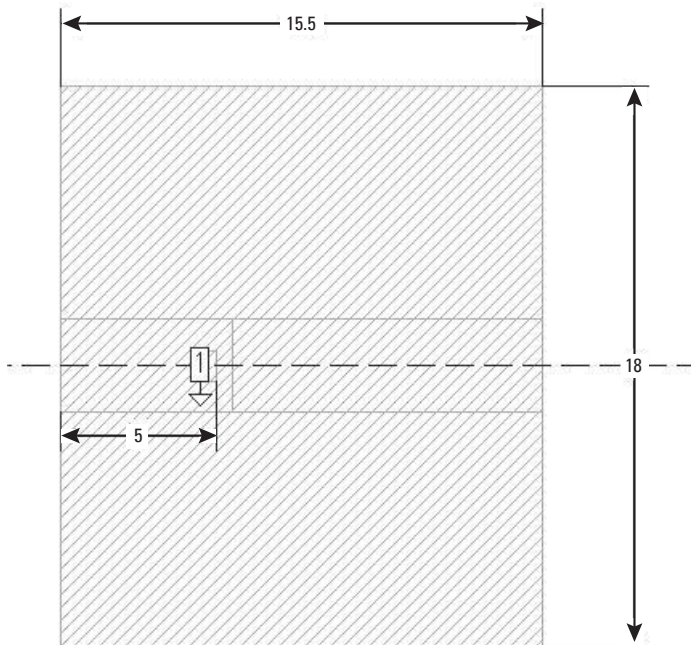


Figure 7.9 Patch antenna with the dielectric substrate increased to 2 mm thick.

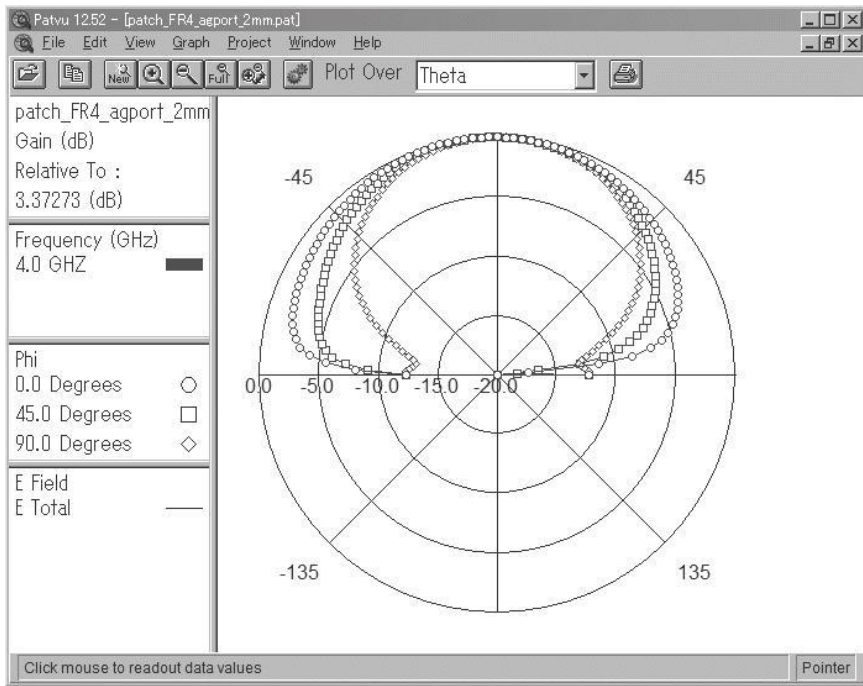


Figure 7.10 Radiation pattern of patch antenna on a 2-mm substrate at 4 GHz. Radiation efficiency is 46%.

7.2.1 Definition of Antenna Gain

Maxwell's equations can be used to precisely calculate that a one-half wavelength dipole has 2.15 dBi gain. As we have mentioned, what this means is that 1 W into a dipole generates a signal in a distant receiver that is 2.15 dB stronger than that same 1 W into a theoretical isotropic antenna. This assumes that the dipole is oriented (broadside) for maximum received signal.

To more firmly understand gain, we can think of the gain of an operational amplifier (op-amp). In this case, the total signal power coming out is stronger than that which went in. The extra signal power comes from a DC power supply being used to enhance the input signal. However, an antenna has no extra DC power supply. So how can it have gain? Even though a dipole has 2.15 dBi of gain, it does not output 1.64 W of electromagnetic waves when we feed it with just one watt. Rather, an antenna with gain outputs more power in one direction and less power in another. If we look at the total power radiated in all directions, no extra power is created.

To further illustrate that dBi is the gain with respect to an ideal isotropic antenna, which radiates power equally well in all directions, see Figure 7.11.

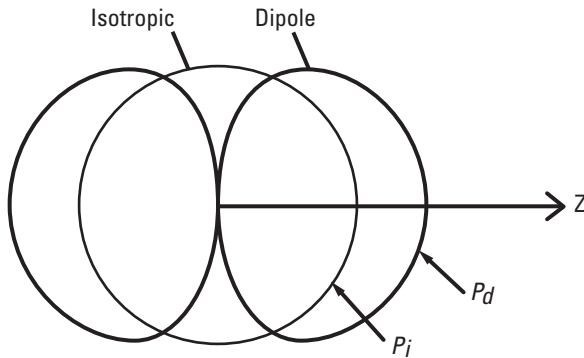


Figure 7.11 Radiation pattern of an isotropic antenna and a dipole antenna.

When P_d is the radiated power of a half wave dipole, and P_i is the radiated power of an isotropic antenna, the gain is the ratio of P_d to P_i in dB. When the gain is relative to an isotropic antenna, as it is here, then we use the units of dBi. Notice that the gain depends on direction. When the direction is not specified, then it is usually the direction of maximum radiation. Just keep in mind that this is not always the case.

The radiation pattern of a three-element Yagi antenna (Figure 7.12) has a higher maximum gain than a dipole. As compared to a dipole antenna, it has maximum radiation in one direction (this is called the forward direction). In a plot like this, the gain might not be scaled with respect to an isotropic antenna. It could instead be scaled with respect to the maximum gain of a dipole antenna. To indicate this, we list the units as dBd.

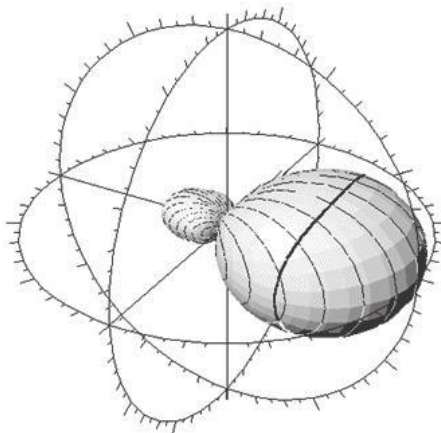


Figure 7.12 Radiation pattern of a three element Yagi antenna.

A dipole has a maximum gain of 2.15 dB relative to an isotropic antenna. Thus, if gain is specified in dBi, just subtract 2.15 to get dBd. If gain is listed in dBd, just add 2.15 to get dBi. Note that the gain in dBi is always larger. If you see an advertisement for the gain of an antenna and it just says dB, then it is probably dBi, because dBi is the bigger number.

In Japan, gain is defined as “the ratio of the electric power fed at the input of a given antenna to electric power required at the input of a reference antenna to generate the same strength of electric field at the same distance in the given direction,” according to Japanese Radiation Law regulation 74, Article 2.

In the case of a simulation, the calculated gain is called the directive gain (G_d) when it is the ratio of the electric power density in the specific direction to the radiated power averaged over all directions. As for a lossless antenna, $G_d = G_a$. When loss is added to an antenna, G_a decreases, but G_d stays about the same.

7.2.2 What Is the Actual Gain?

Figure 7.13 is a simulated result of a one-half wavelength dipole. When the surface resistance of the thin rod forming the dipole is $0.05 \Omega/\square$ (read, “ohms

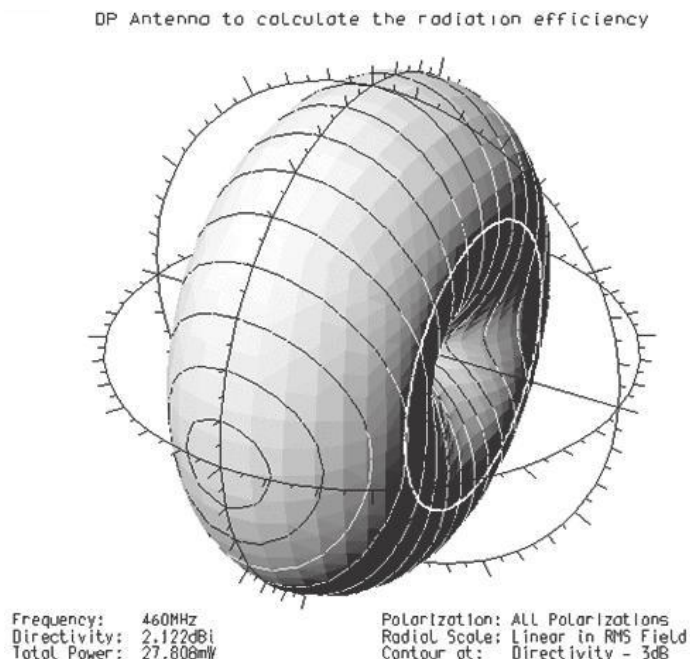


Figure 7.13 Simulated result of the one-half wavelength dipole.

per square”) P_{rad} is 27.8 mW, and P_{loss} is 0.5 mW at 460 MHz, the radiation efficiency η is calculated as follows.

$$100 \times P_{rad} / (P_{rad} + P_{loss}) = 97.9\%$$

Here, the actual gain is reduced by the less than perfect radiation efficiency that is caused by antenna loss. For this example, a dipole, the directive gain G_d is 2.12 dBi:

$$\begin{aligned} \text{Actual gain in dB} &= \text{Directive gain in dB} + 10 \log_{10} (P_{rad} / (P_{rad} + P_{loss})) \\ &= 2.12 - 0.09 = 2.03 \text{ dB} \end{aligned}$$

The actual gain is the gain we use to calculate the actual signal strength at the receiver.

7.2.3 Measuring the Antenna Gain

When we actually manufacture the antenna and we would like to measure the gain, something resembling the method shown in Figure 7.14 is often used. For this test setup, a signal generator on the left plays the role of a transmitter and the electric field strength meter stands in for the receiver. A high-quality attenuator is important to minimize any mismatch between the antenna and the receiver. The attenuator is also used to adjust received signal strength.

The antenna on the right, connected to the field strength meter, is the antenna under test. It should be located at least a few wavelengths from the transmitter so that the antennas are each in the far field of the other. Higher gain antennas must be positioned farther apart.

We must also make sure there are not any extraneous objects in the way. For example, Hertz initially measured a wildly incorrect value for the speed

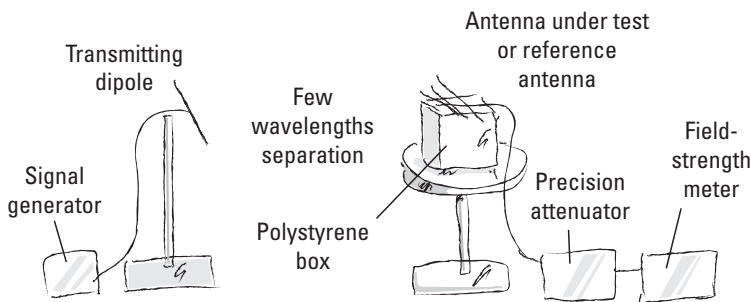


Figure 7.14 Measuring antenna gain using a signal generator, attenuator, and field-strength meter.

of light in one of his early experiments. It turns out that the problem was a large cast iron stove located nearby in his laboratory. Effects of objects of any kind, including walls, earth, concrete, people, and so forth, can seriously disrupt measurements. Use of an anechoic chamber is expensive, but it does allow very high-quality measurements. An anechoic chamber has its walls completely covered with RF-absorbing material.

To proceed with a measurement:

1. Connect a half-wave dipole to the receiver and measure the signal strength. This dipole is the reference dipole. Our measurements will be referenced to this dipole, thus we will generate measurements in dBd. Since isotropic antennas exist only on computers; most physical measurements are relative to a reference dipole.
2. Set the attenuator so the receiver indicates a comfortable value of signal. Too little signal and we have increased the noise. Too much signal and we risk pushing the receiver into a nonlinear response and our measurements will be useless.
3. Take out the reference dipole and connect the antenna under test. Measure the received signal strength. Rotate the antenna for maximum received strength. If the direction seems odd, look for extraneous objects that are confusing the antenna radiation pattern.
4. Adjust the attenuator so that the received signal strength is the same as it was for the reference dipole. The change in the attenuator value is the value of the actual gain, dBd.

7.2.4 Does Higher Gain Mean Higher Performance?

The actual gain indicates the strength of the signal that is received. A higher actual gain means a larger signal. So, a higher gain is always better, right? Many times, the answer is yes. But not always. For example, if both the transmitter and receiver are not moving, and we can take the time to point their antennas for best signal strength, then higher gain can certainly help. It not only gives our desired receiver a better signal, but other, undesired receivers receive a weaker signal. On top of that, we attenuate the signals from other undesired transmitters.

In another situation, a large number of PCs might want to communicate with a central access point such as a wireless LAN. A wireless LAN antenna with a highly directive antenna will have many dead zones where PCs cannot

connect. In these applications, it is good for the gain to be close to the 0 dBi of an isotropic antenna. Fortunately, for this application, most PCs will not be above or below the wireless LAN, so the wireless LAN can concentrate its signal equally in all horizontal directions, all the while reducing signal upwards and downwards. This kind of antenna is referred to as omnidirectional or an omni, for short, even though power is mostly radiated only in horizontal directions.

Thus, it is hard to say that the higher the gain, the better the performance. We should evaluate and try to achieve directive gain and actual gain depending on the application.

To see what a lossy environment can do to an antenna, Figure 7.15 shows the gain of the patch antenna with $\tan\delta = 0.05$, double the amount we used for the same model as shown in Figure 7.7. The gain, listed in the upper left, is negative, -0.56 dBi. With the lower loss substrate (Figure 7.7), we had positive 1.7 dBi gain. Thus, typically in the case of small antennas in a complex, lossy environment such as a PC, higher gain indeed usually indicates a better antenna.

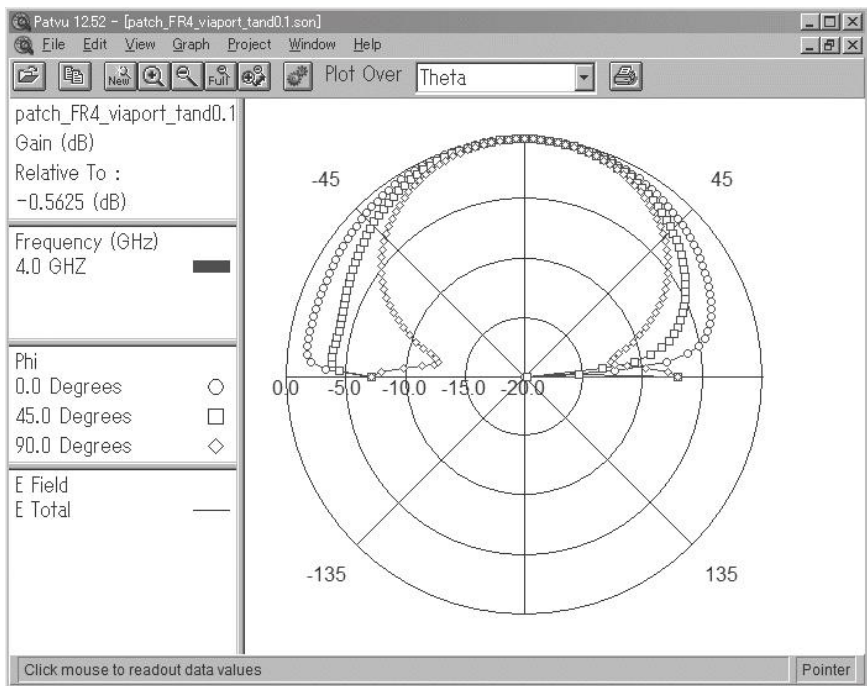


Figure 7.15 Gain of a patch antenna with dielectric loss doubled to $\tan\delta = 0.05$.

7.2.5 How a Reflector Influences Directivity

Figure 7.16 is the result of a dipole antenna for 2.45 GHz that is one-quarter wavelength (3.1 cm) in front of a metallic wall. The electric field strength distribution is represented.

The left wall is made of an ideal conductor (a so-called electric wall). The other five surfaces of the solution space are defined to be absorbing boundaries. We are viewing a vertical dipole antenna from above. The displayed plane cuts the center of the vertical element and displays the electric field distribution. The actual EM wave oscillates sinusoidally with time. The plot shows the root mean square (RMS) value. We see that the dipole has strong radiation in most directions.

Next, Figure 7.17 shows the result of dipole one-half wavelength in front of the wall. Comparing with Figure 7.16, it is clear this dipole does not radiate any power at all in the direction perpendicular to the wall. That power did not disappear, it is now radiating more strongly in several other directions, which means we now have some directivity. The change in distance from the wall is only one-quarter of a wavelength. What is happening?

The difference between these two field distributions may be understood by considering the waves shown in Figure 7.18. The dipole, A , launches two waves, one to the right and one to the left. When the distance between A and R is $\lambda/4$, the leftward launched wave is delayed by 90° on its way to reflection at R . Upon reflection, the wave is flipped 180° . Notice that this means that the total

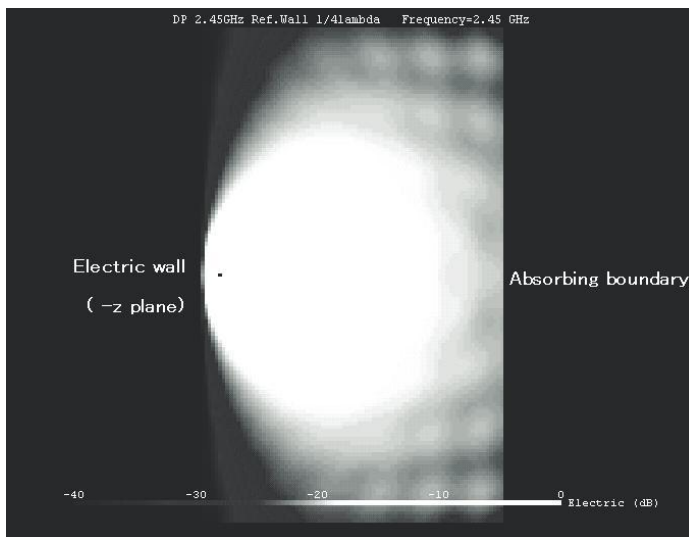


Figure 7.16 Electric field strength distribution for a dipole antenna one-quarter wavelength (3.1 cm) in front of a metallic wall.

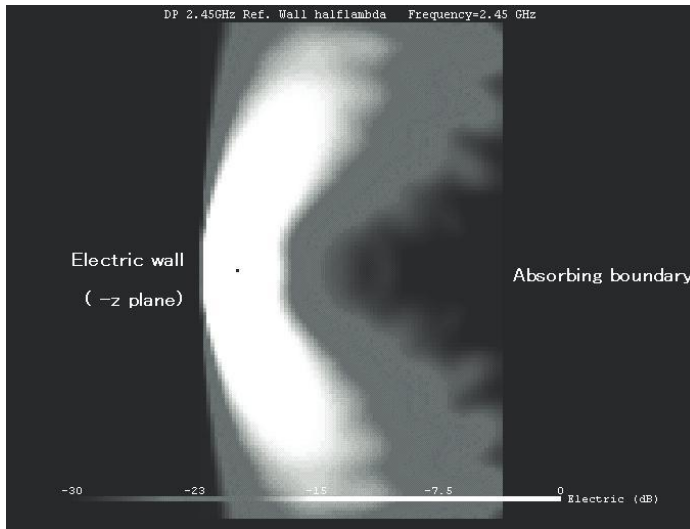


Figure 7.17 Electric field strength distribution with a dipole antenna in front of a metallic wall by a distance of one-half of a wavelength.

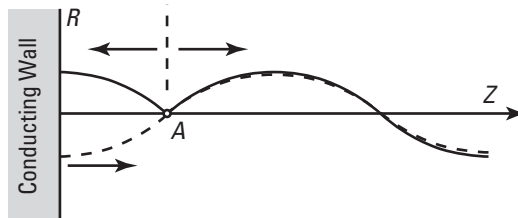


Figure 7.18 Fields around a dipole, A , positioned $\frac{1}{4}\lambda$ from a reflecting wall, R .

electric field (summing up both incident and reflected waves) is zero. Zero total electric field on a conductor is called a *boundary condition*. Then, on the way back to the dipole, it is delayed another 90° . The grand total is 360° . As far as phase is considered, that is just like it is not delayed at all! So, the reflected wave is exactly in phase with the wave that the dipole launches to the right. They add constructively to give maximum radiation to the right.

When the dipole is $\lambda/2$ from the wall, we can use the same line of reasoning. Now, the reflected wave is delayed $180^\circ + 180^\circ + 180^\circ = 540^\circ$ before it gets back to the dipole. This is the same as 180° . Now the reflected wave is exactly out of phase with the transmitted wave and they cancel. There is no radiation to the right.

Access point equipment such as wireless LANs often use dipole antennas. If we set up such an antenna near a large, flat metal wall, we should take care that the distance is not $\lambda/2$ as it is in Figure 7.17. Keep in mind that this can be a problem whenever the distance is an integral multiple of a half wavelength, as this gives us a similar situation. Fortunately, an object that is farther away must also be much larger in order to reflect the same amount of power. When the distance becomes large enough, there are plenty of small random factors that come into play.

7.2.6 Standing Waves Between Two Metal Walls

Figure 7.19 shows the electric field for a dipole between two conducting walls. The dipole is $\frac{1}{4}\lambda$ from the left wall. Another electric wall is located far to the right. The waves radiated from the dipole in both directions bounce off of both walls, back and forth. It is actually a large echo chamber.

Recall that when we have two waves traveling in opposite directions, we have standing waves. We can see these stationary standing waves in many places. If we position a receiving antenna at one of the standing wave peaks, we get a strong signal. If, on the other hand, our receiver is at one of the nulls, we receive no signal at all. To counter this problem, you might notice that some PCs and some wireless LAN routers have two antennas. If one antenna is in a null and receiving no signal, they just switch to the other antenna. This is called space diversity.

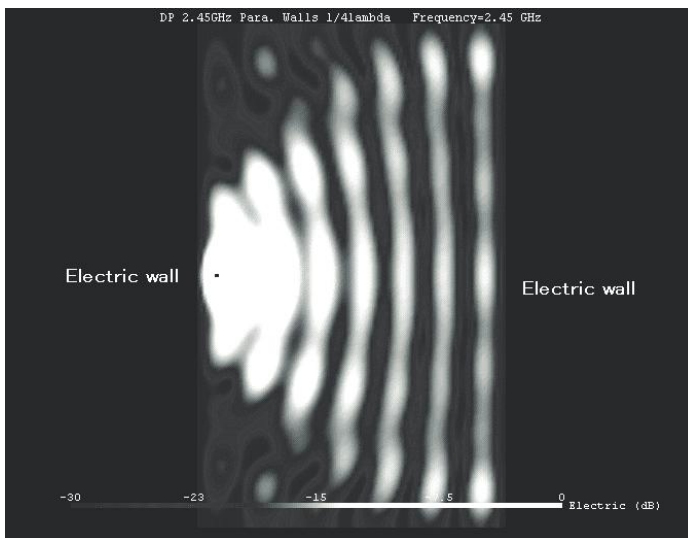


Figure 7.19 Electric field from a dipole $\frac{1}{4}\lambda$ from the conducting wall to the left, and a similar conducting wall far to the right.

Another problem with multiple signals bouncing around is called multipath. You might remember this from “ghosting” on the old style televisions. In the case of wireless LANs, if the reflected signal is delayed by, say, the time required to send one data bit, the receiver is slapped with two different data bits at the same time. This problem can be addressed with various coding and modulation schemes, but that is outside of our topic.

7.2.7 Magnetic Current Antennas

The nulls in Figures 7.17 and 7.19 are nulls in the electric field. Since we are using an electric field detection type of antenna (a dipole), the received signal goes to zero whenever the electric field goes to zero. What would work in this situation would be a dipole that uses magnetic current instead of electric current. Magnetic current is the flow of the magnetic equivalent of the electron—a magnetic monopole. Unfortunately, while magnetic current is allowed in Maxwell’s equations on equal status with electric current, we do not see magnetic current anywhere in nature. So much for that idea.

Actually, it turns out that we can make something that uses electric current but behaves just as if it were a magnetic dipole. Remember the magnetic field detection type of antenna from Chapter 1? Just a small coil of wire, perhaps wrapped on a ferrite rod to increase the magnetic sensitivity. It makes sense. In a standing wave when we have minimum electric field, we have maximum magnetic field. So if the electric field antenna fails, simply switch to a magnetic field antenna.

7.3 Bandwidth of Antennas

Here is a great idea: Let’s get a substrate with a really high dielectric constant (or magnetic permeability) and build a really tiny antenna on it. In principle, no problem. In practice, a couple things get in the way. One of them is bandwidth.

7.3.1 Definition of the Bandwidth

As we have seen in previous chapters, we like the reflection coefficient, S_{11} , seen at the feed point of an antenna (port 1 for electromagnetic field simulators) to be as small (i.e., large negative dB) as possible. This gives us a good match to the transmitter and receiver and allows a strong signal to be formed.

Figure 7.20 shows the reflection coefficient for several half-wave dipoles. So, what is the bandwidth? The precise definition is up to the designer. For example, we could say that the bandwidth is the frequency range for which the reflection coefficient is under -10 dB. Someone else might say -6 dB. The

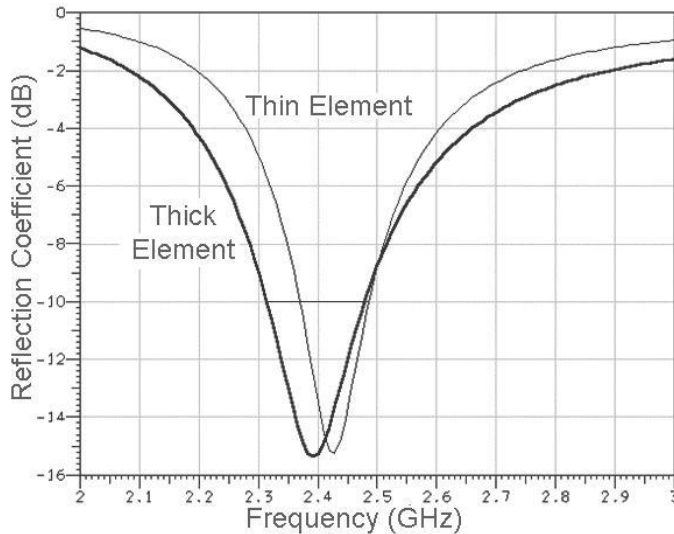


Figure 7.20 Bandwidth is the frequency range over which the reflection coefficient is under -10 dB ... or maybe -6 dB. It is our choice.

important thing is that everyone who needs to communicate on a given project agrees on the same definition.

In this simulation, we compare two antennas, one uses a thinner wire $10\text{ }\mu\text{m}$ in radius and the other uses a thicker wire $100\text{ }\mu\text{m}$ in radius. Both are 60 mm long. We notice that the antenna with thicker wire has wider bandwidth. This should sound familiar. The broadband bow tie antenna of Section 5.3 can be thought of as using thicker wire.

The voltage standing wave ratio is the ratio of the peak standing wave voltage on a transmission line to the minimum standing wave voltage. If the line is perfectly matched, then the magnitude of the reflection coefficient is zero and the VSWR is 1. This means the voltage on the transmission line is the same everywhere. There is no standing wave.

If the transmission line is terminated in an open or short circuit, everything is reflected, the magnitude of the reflection coefficient is 1, and the VSWR is infinite. This is because the minimum transmission line voltage is zero, and we must divide by zero to calculate VSWR. This means the VSWR is infinity.

In general, the VSWR is given by the formula shown below.

$$VSWR = (1 + |S_{11}|) / (1 - |S_{11}|)$$

Figure 7.21 indicates where the VSWR is 2, or as is also commonly written, 2:1, and stated as two to one. This level corresponds to a reflection coefficient of about -10 dB. This is a frequently used definition of bandwidth.

Resonant antennas like the dipole have narrow bandwidth and thus relatively high Q . The relationship between the bandwidth, B , and Q is given below.

$$B = f_0 / Q$$

The bandwidth, B , divided by f_0 , is called the fractional bandwidth, is just one over Q .

7.3.2 Design of a Wideband Dipole Antenna

Figure 7.22 shows a dipole 240 mm long that we simulated in Chapter 4. It resonates at around 600 MHz (Figure 7.23).

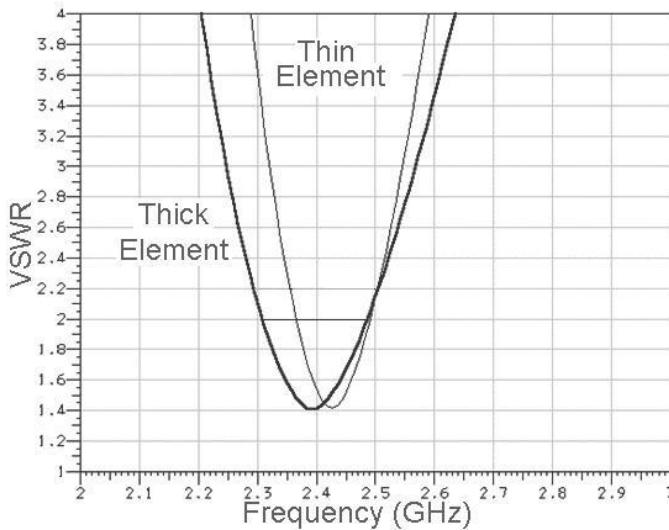


Figure 7.21 VSWR = 2 corresponds to return loss of 10 dB.

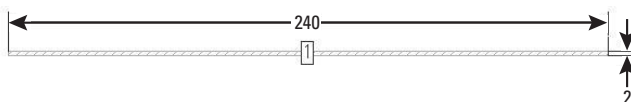


Figure 7.22 Dipole antenna element, 240 mm long, resonates at around 600 MHz.

Japanese digital terrestrial television channels are allocated between 470 and 700 MHz. If you use this antenna for receiving digital terrestrial television, it works well at the center frequency; however it has a poor return loss (high VSWR) at the band edges, so we need a wider bandwidth antenna.

The typical TV coaxial cable has a characteristic impedance of 75Ω , so we need to evaluate the reflection coefficient normalized to 75Ω . As the R of the input impedance of a one-half λ dipole antenna is almost 75Ω , it is suitable for a 75Ω coaxial cable.

Just to see what will happen, we will work this problem differently from what we did previously. Let's increase the line width. Figure 7.24 shows the dipole element increased to 12 mm wide, with the length unchanged at 240 mm.

As for the ultrawide band antennas, we can back off our bandwidth definition to a return loss of better than 6 dB (this is a VSWR of 3:1). In this case, this antenna almost achieves the required bandwidth (Figure 2.25).

Maybe wider is better. However, it might be nice to reduce the amount of metal. So, we can make the dipole into a kind of net. We put more conductor

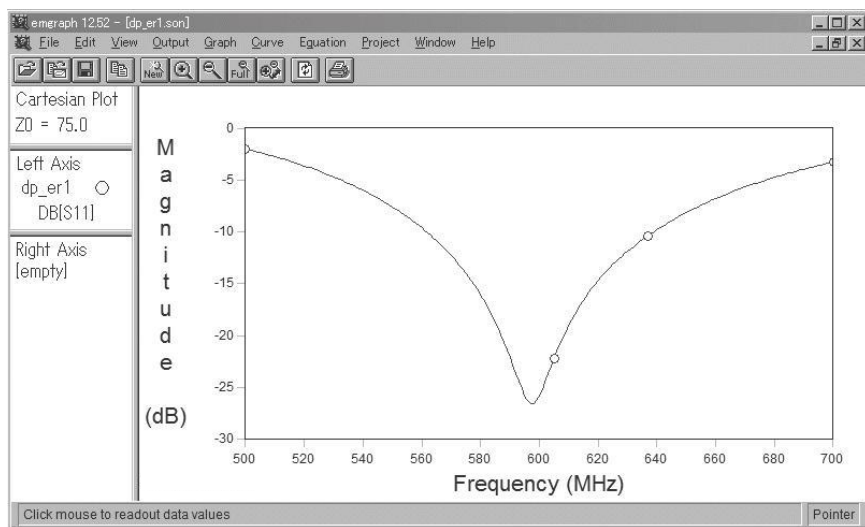


Figure 7.23 Reflection coefficient of the dipole antenna normalized to 75Ω .

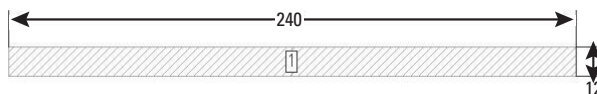


Figure 7.24 Dipole line width increased to 12 mm.

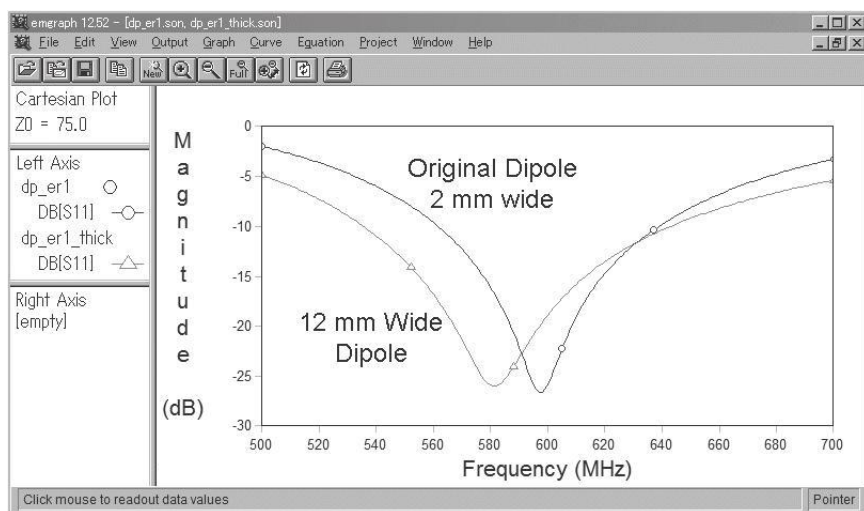


Figure 7.25 Reflection coefficient when the dipole line width is increased to 12 mm.

where there is more current, which is always near the edge of planar lines (Figure 7.26).

The reflection coefficient of the antenna (Figure 7.27) now shows an even wider bandwidth.

7.3.3 A Wideband Patch Antenna

We have learned about the patch antenna in Chapter 4. The bandwidth of an antenna is determined by the structure, dimensions, materials, and any adjustable parameters such as the thickness and the dielectric constant of the substrate. For example, by simulating various thicknesses, we can see that the bandwidth changes slightly. However, we can only make a substrate just so thick before we try to find easier ways to expand the bandwidth.

A triple-patch antenna (Figure 7.28) included in the Sonnet examples (Help->Examples->Antennas, File: tripat.son) and developed by Matra Defense,

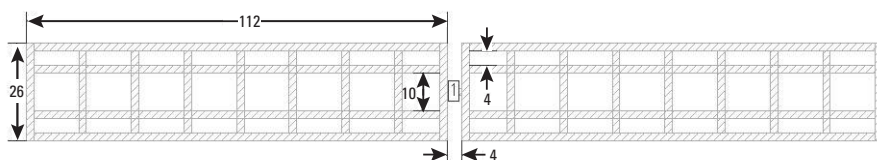


Figure 7.26 A netted structure gives us the broadband effect of a wide-width dipole without using a lot of metal.

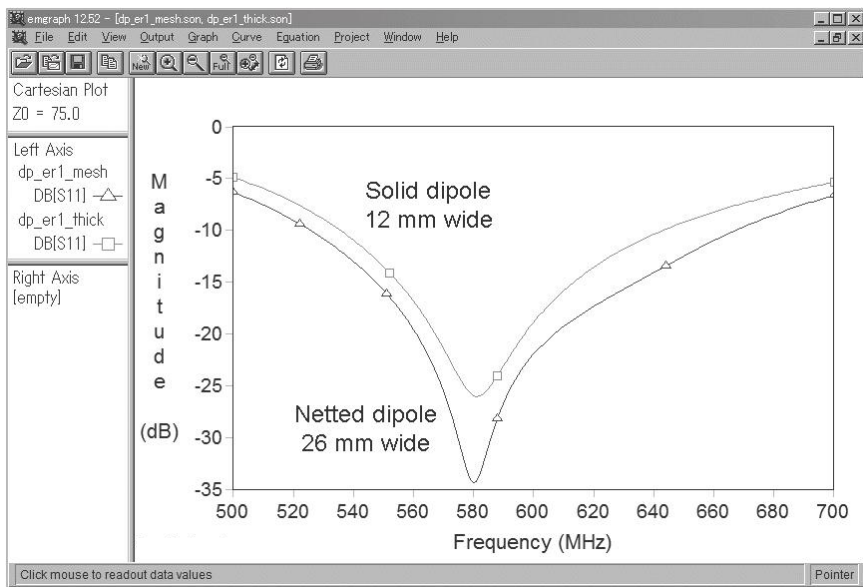


Figure 7.27 Reflection coefficient of the wide line width, netted dipole (triangles) compared to the 12-mm-wide line width dipole (rectangles).

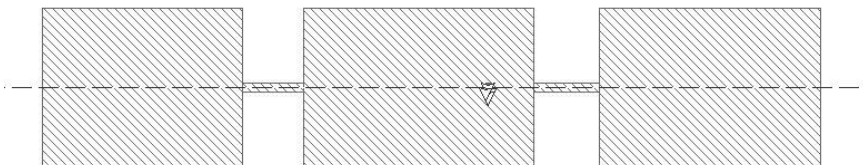


Figure 7.28 Triple-patch antenna developed by Matra Defense, Antennas & Stealthiness Dept. in France, included in Sonnet examples (File: tripat.son).

Antennas & Stealthiness Dept. in France, is one way to make a broadband antenna.

The dimensions of each patch are 36×28 mm, 41.9×28 mm, and 39.9×28 mm from the left, the relative permittivity of the substrate is 2.94, and it is 3.04 mm thick. Figure 7.29 is the VSWR of this antenna, the thin line shows the result of Sonnet, and the thick line is the result of measurement. The difference between the measured and calculated resonant frequencies is about 1%. This is typically considered very good agreement for this type of problem.

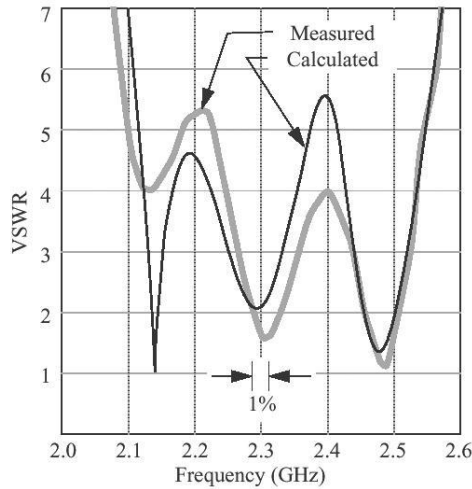


Figure 7.29 VSWR of the triple-patch antenna.

As this antenna gives a fairly high VSWR, this specific design might be best used for three different bands. However, with careful tuning of the individual patch resonant frequencies, it should be possible to obtain a good broad band response.

7.3.4 Wideband Double Patch

The triple-patch antenna connects all three patches with feed lines. All three patches being next to each other makes the complete antenna longer and requires more space to mount. So, we will try a double-patch antenna, with one patch above the other.

Figure 7.30 shows a model using this idea; both patches are 15.5×18 mm each, and the upper one is not fed directly. The thickness of dielectrics are 2 mm each, with a relative permittivity of 4.9, and $\tan\delta$ is set to 0.025.

We want to tune the location of the port for best return loss. Therefore we need to set the location of the port as a parameter (Figure 7.31). First, click on Add Points button (shown in the figure) in the upper right corner of the Tool-Box. Then click on the edge of the patch to add a point that will become the anchor point for our parameter (indicated by the white square in the figure).

Next, select Tools->Add Dimension Parameter->Anchored, and click on the point we just added, and press Enter. Then, follow the next instructions. We assume you have already added parameters to previous Sonnet projects. If not, you might want to review the previous chapters where we did so, or check

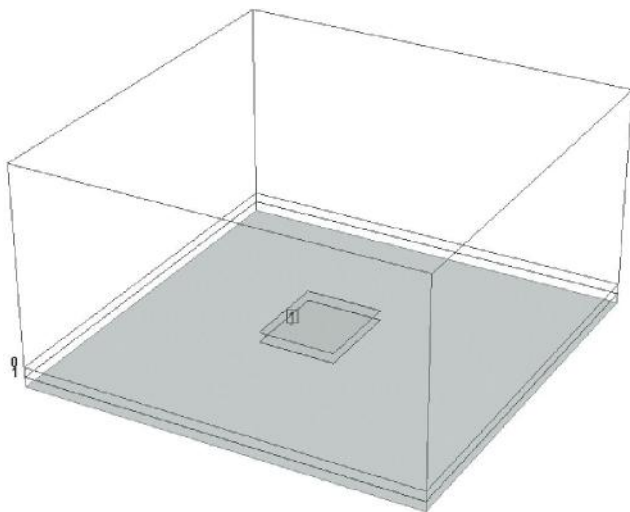


Figure 7.30 Both patches are 15.5×18 mm each, and the upper one is not fed directly.

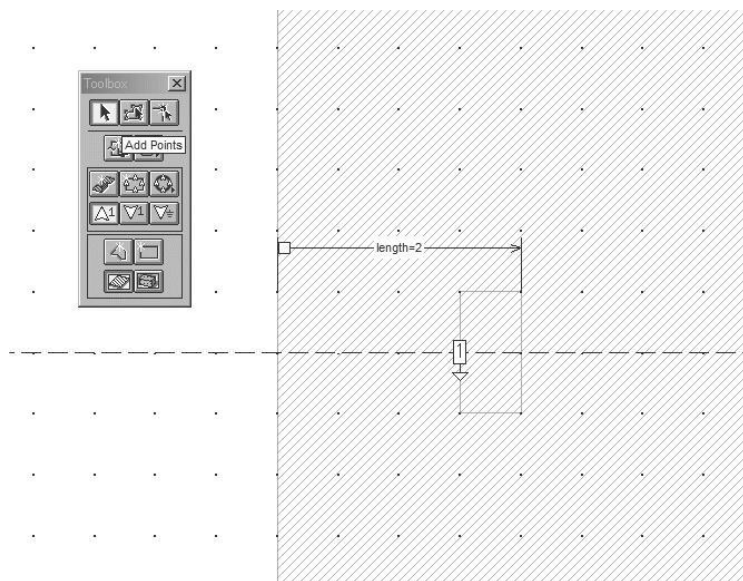


Figure 7.31 Parameter used to adjust the port position. File: patch_FR4_agport_2mm_2.son.

the Sonnet documentation under Help. Note that we are adding an “Anchored” parameter, rather than the “Symmetric” parameter that we did earlier.

1. Click on the top right corner of the two-cell rectangle to which the autogrounded port is attached to display small white squares.
2. Drag to select the area including all four corners of the two-cell rectangle used for the port.
3. The four corners are now displayed as small black squares. Press Enter and then input a name for the parameter on the displayed dialog box.
4. Check to make sure you have specified the parameter correctly by double-clicking on the parameter name and typing in a new value for the parameter.

To set up our analysis, select Analysis->Setup and chose Parameter Sweep under Analysis Control and press the Add button. The dialog box shown in Figure 7.32 is displayed. Then check the Length parameter check box and specify a sweep from 1 to 5 with a step of 2 mm. Then select Project->Analyze.

As we have done for the other parameterization examples, double-click on DB[S11] in the plot, check Graph All Iterations in the bottom left of Parameter Combinations, and click OK. All reflection coefficient curves for all values of the parameter are now displayed (Figure 7.33).

Two frequencies where the reflection coefficient is small are found, with a somewhat higher reflection in between. In these results, the case of length = 1.0 mm (highlighted) seems like a good wideband compromise.

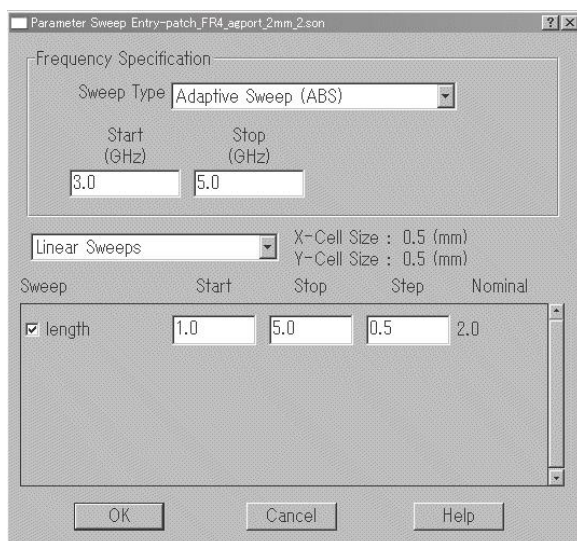


Figure 7.32 Set 3 GHz for Start and 5 GHz for Stop by Adaptive Sweep (ABS).

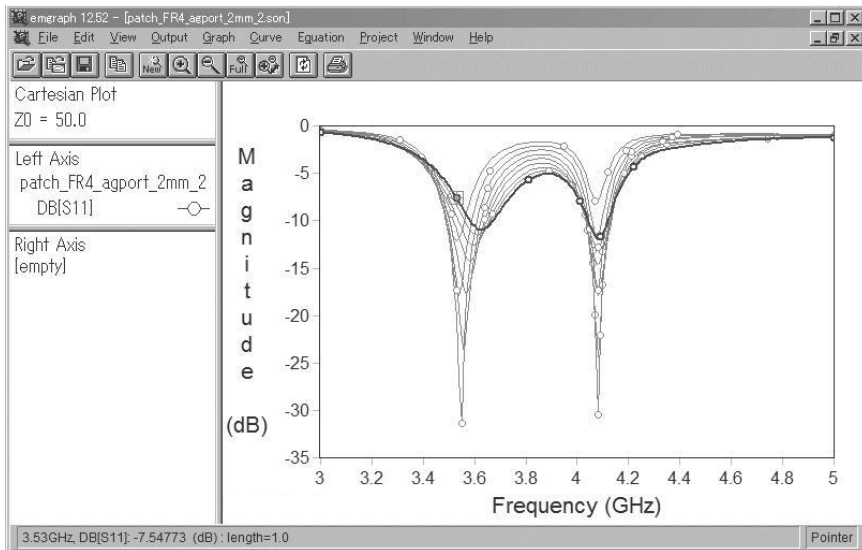


Figure 7.33 Reflection coefficient for all values of the parameter.

Two resonating objects (in this case, two patches), which are close enough to strongly couple and whose resonant frequencies are very close (or even identical) to each other, tend to show two resonances that have been pulled to lower and higher frequencies. This gives the characteristic W-shaped graph we see here.

This “bimodal characteristic” can be adjusted by changing the distance (i.e., coupling) between the resonators. This tuning can be very sensitive in some cases. Electromagnetic simulators are especially useful for optimizing the distance and any other dimensions. If we were to tune by repeatedly prototyping the antenna, the job can become most time- and budget-consuming.

7.3.5 Parallel Configuration of Patch Antennas

Is it possible to get wider bandwidth by placing two patches side-by-side on the same plane? In this case, we cannot use the Symmetry feature of Sonnet, so we model a pair of rectangular patches without symmetry turned on (Figure 7.34). The second patch, the one without a feed, is shorter.

The port location is fixed at 4 mm from the edge of the patch, and the length of the patch without a feed is set to be a parameter. Symmetrical change of the patch length is set by selecting Tools->Add Dimension Parameters->Add Symmetric, just like the dipole antenna in Chapter 3.

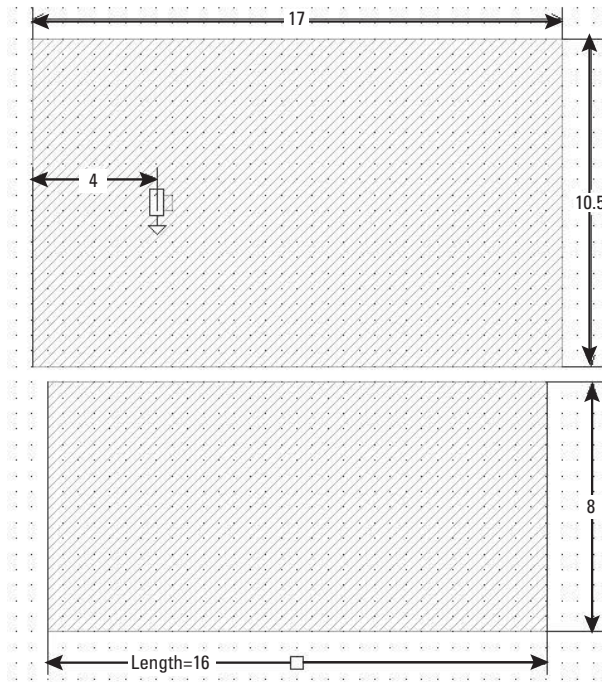


Figure 7.34 Dual, side-by-side rectangular patches; the length of the lower patch is set to be the parameter. File: patch_FR4_agport_2mm_para.son.

The reflection coefficient (Figure 7.35) is shown for all analyzed values of the parameter. The highlighted curve is for a separation of 17 mm and is a nice compromise for wideband operation. Notice that we have the same characteristic W curve. The surface current distribution for this case is shown in Figure 7.36.

Animations of current distributions of multipatch antennas can be quite beautiful. Select Project->View Current. Then select Animation->Settings and check the Time radio button (Figure 7.37). Click OK, and next select Animation->Animate View to see the time animation. Use the VCR controls that appear. We see how the current at 4.06 GHz changes as we progress through an RF sine wave excitation applied to port 1. Note when the current on the lower patch flows most strongly.

At some frequencies, both patches radiate power. However, the radiated power might not be in phase, so the radiation pattern and radiation efficiency should be carefully checked. Figure 7.38 shows the far field radiation pattern at 4.06 GHz. Because both patches are radiating, the radiation pattern is slightly

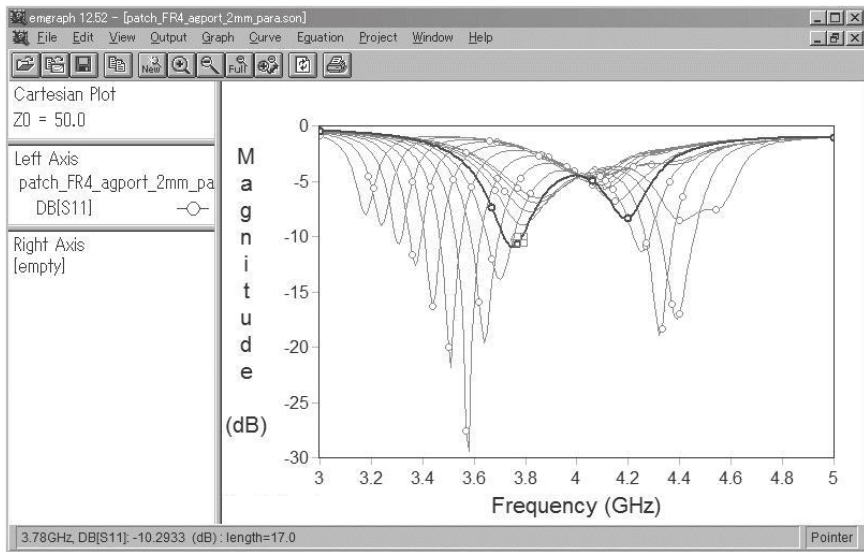


Figure 7.35 Reflection coefficient for all analyzed values of the parameter.

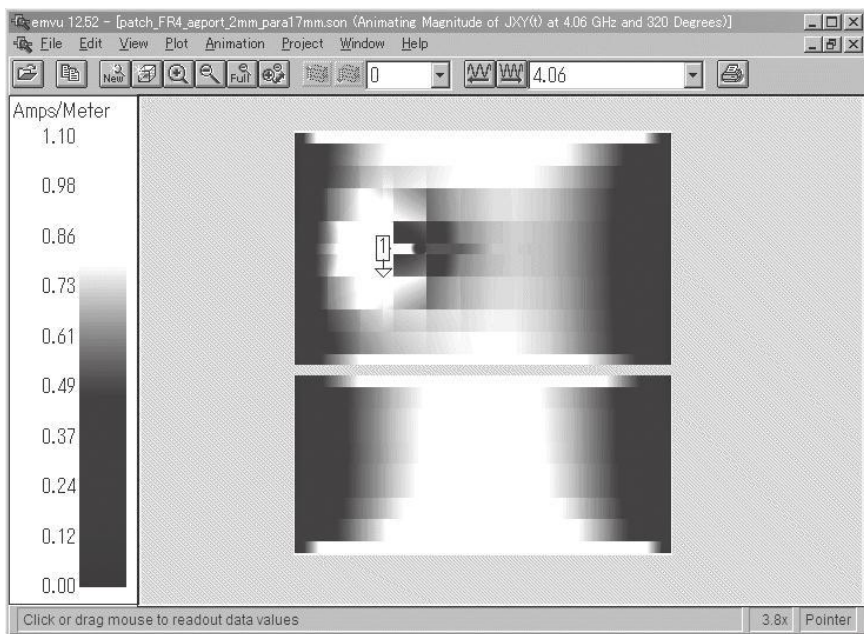


Figure 7.36 Surface current distribution on the patch for length = 17 mm at 4.06 GHz.

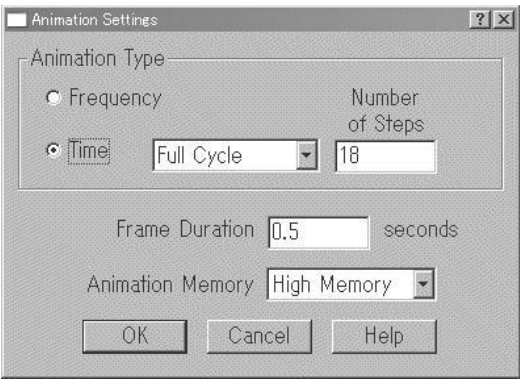


Figure 7.37 Current distribution animation settings.

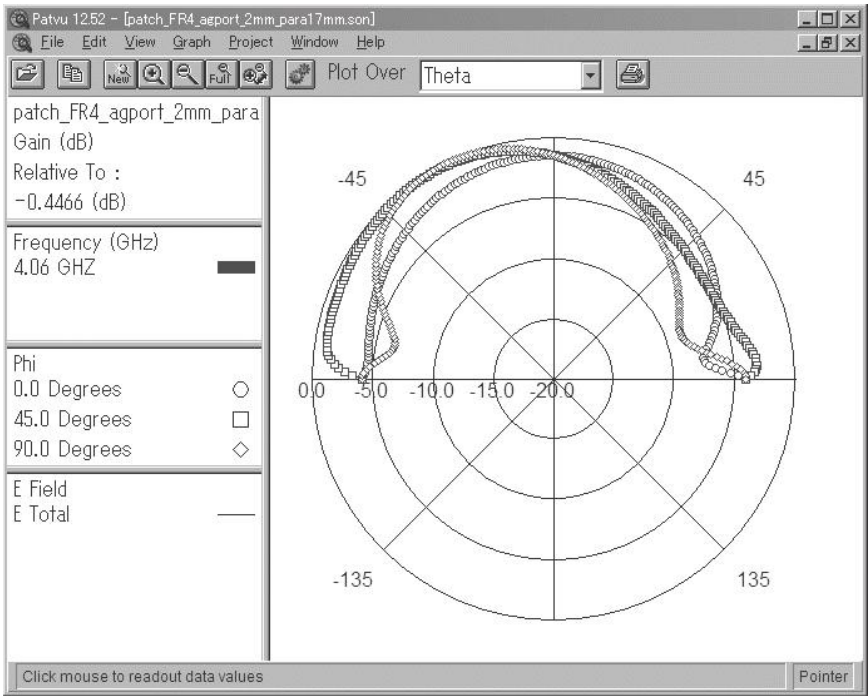


Figure 7.38 Far field radiation pattern is slightly skewed. Directive gain is 6.6 dBi at 4.06 GHz, and the actual gain is -0.45 dBi.

skewed. The directive gain is 6.6 dBi at 4 GHz; however the actual gain is -0.45 dBi, so the radiation efficiency dropped to 20%.

This patch antenna has a dielectric thickness of only 2 mm. A $\tan\delta$ of 0.025 means the loss is a bit high. One way to improve the radiation efficiency is to reduce the dielectric loss. Also, the substrate thickness should be kept as large as possible, as it is difficult to radiate from thin substrates.

Therefore, we must get our patch antenna as far as possible from its ground. Given that the patch antenna is fed by a via, we must be careful when the substrate thickness approaches a one-quarter wavelength. Our via feed might start to behave like a monopole antenna and its radiation pattern will start to interfere with our patch radiation pattern. In addition, substrate thickness affects wavelength shortening.

Instead of thin dielectric substrates, a patch can be formed from a self-supporting metal plate mounted on a wall or conducting surface with spacers. Now there exists only low-loss air between the patch and its ground, so there is no dielectric loss and radiation efficiency is improved. Keep in mind that we still have the resistive loss in the conducting plate that reduces radiation efficiency.

7.3.6 Wideband Short-Circuit Patch Antenna

It would be really nice to miniaturize a single-patch patch antenna and to make it wideband at the same time. Figure 7.39 shows one such possible antenna with two vias on both sides shorted to ground.

For the fundamental (one-half wavelength) mode, the center line along the width of a rectangular patch has zero electric field between the patch and ground. So, it should not affect anything to place vias anywhere along this line.

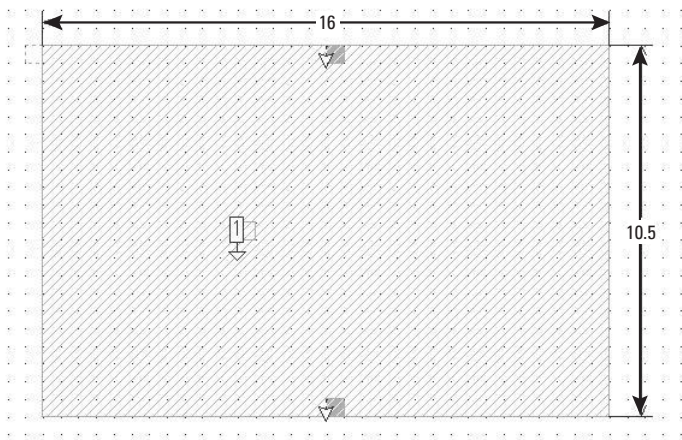


Figure 7.39 Model with two vias on both sides to be shorted to ground. File: patch_FR4_agport_2mm_short.son.

Figure 7.40 shows the reflection coefficient, and it is found to be matched to 50Ω at 4 GHz.

The surface current distribution on the patch is observed to be the expected one-half sine wave along the edge, even with the shorts in place (Figure 7.41). However, if a shorting via moves from the center line, it causes a change in the current distribution. The affected half-wavelength current distribution loses its sine wave symmetry.

To use this idea, we can modify the effective length of current flowing along either edge in order to achieve our design goals. Figure 7.42 the parameterized model in Sonnet that allows moving the upper via along the edge by means of a parameter. Select Tools->Add Dimension Parameter->Anchored to define the length from the edge of the patch to the via. Draw a one-cell square rectangle just outside the top left corner (Figure 7.42) to which we can anchor the anchored parameter. When we analyze, we will get a warning message about polygons outside of the box. Just ignore that message. Use a similar rectangle on the ground as a base for the via.

Figure 7.43 is the reflection coefficient for all values of the parameter. Just as we found for the two double-patch antennas, the reflection coefficient has the characteristic W shape.

The input impedance is well matched to 50Ω and the resonant frequency shifted slightly higher. In Figure 7.42, the shorting position is at length = 4 mm, and the now asymmetric surface current distribution on the patch is shown in Figure 7.44.

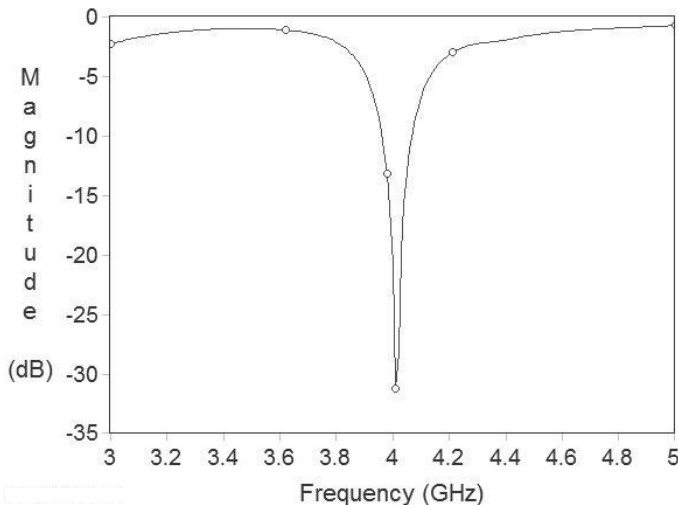


Figure 7.40 Reflection coefficient of a shorted patch antenna, matched to 50Ω at 4 GHz.

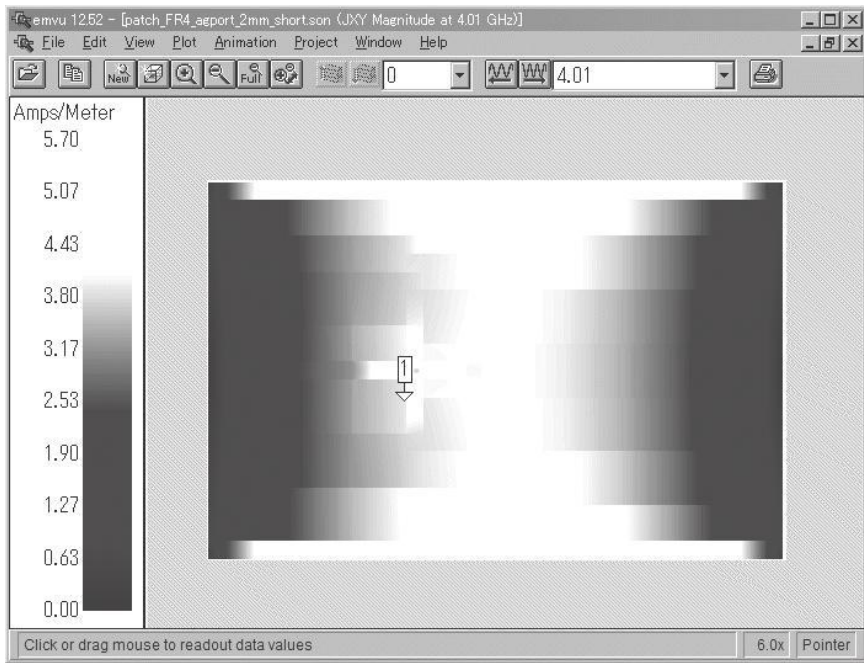


Figure 7.41 Surface current distribution on the shorted patch.

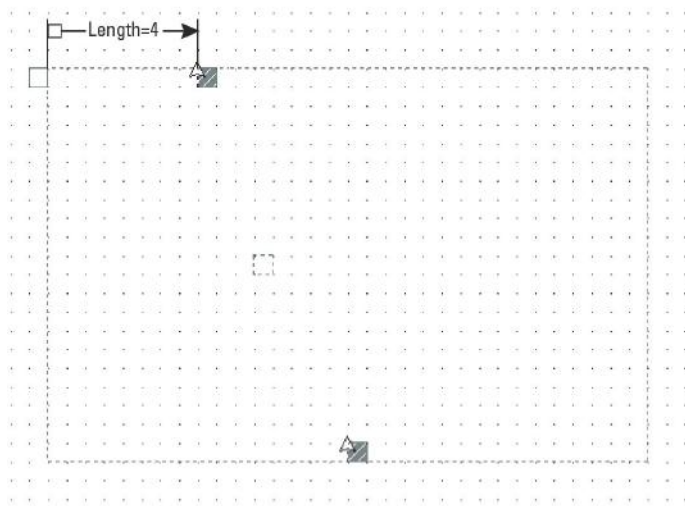


Figure 7.42 Changing the position of one via along the edge by using a parameter. File: patch_FR4_agport_2mm_short.son.

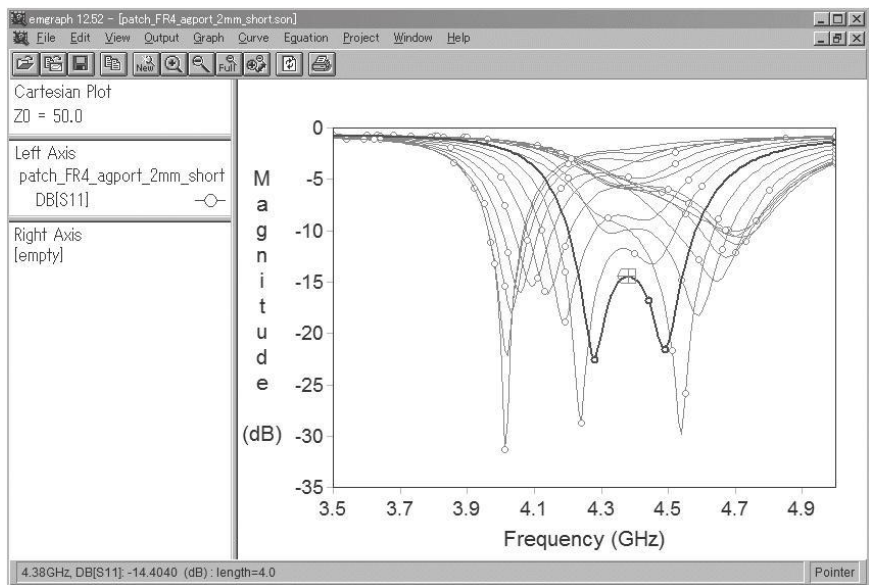


Figure 7.43 Reflection coefficient for all values of the parameter.

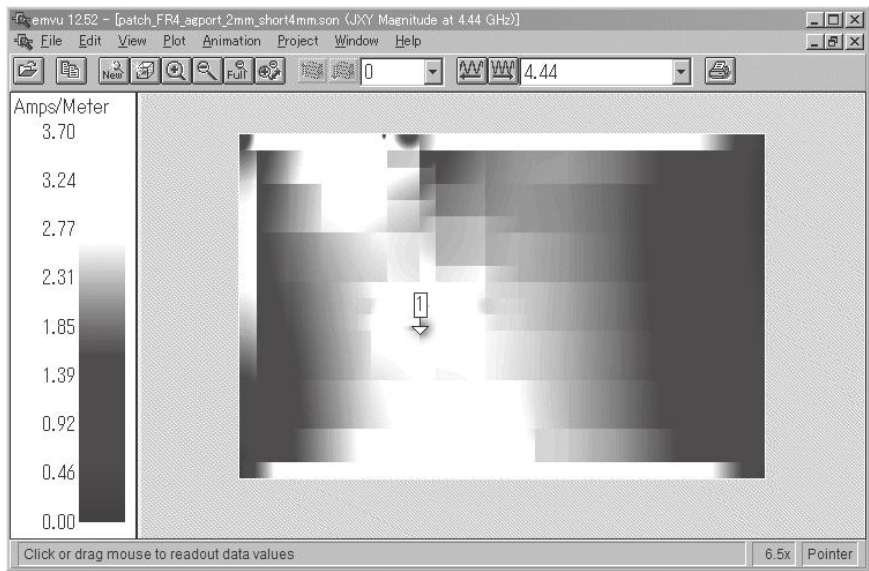


Figure 7.44 Surface current distribution on the shorted patch.

7.4 Interrelation of Three Parameters

In all of engineering, it seems that if we push in on one aspect of a problem, some other aspect pops out. The same is true with antenna design.

7.4.1 Small Antennas and Three Parameters

In the world of small antennas, an empirical rule-of thumb is as follows.

$$\text{Gain} \times \text{Bandwidth/Volume} \approx \text{Constant}$$

For example, if we want higher gain, no problem. But we will likely need extra volume or less bandwidth. A similar empirical rule-of-thumb is as follows.

$$\text{Antenna Volume}/(\text{Gain} \times \text{Efficiency} \times \text{Bandwidth}) \approx \text{Constant}$$

For example, if we want higher efficiency, we will likely need larger volume.

By these rules, when we start with a reasonable antenna design, we can see that it might be difficult to improve all of the parameters, or even to improve one of the parameters without compromising on another parameter. So we should set priorities depending on the application, and decide just how important each parameter really is to the overall success of a project.

8

Practical Antennas

8.1 Ultrawideband Antennas

We have explored lots of different areas in antenna design (and if you are interested, there is a lot more to explore, enough for a lifetime or two!). Let's see how we can put some of this knowledge to work.

8.1.1 What Is a Pulse Excitation?

An electrical pulse is an electric current and voltage of short duration. For example, radio transmission of wireless USB uses ultrawide band (UWB) technology, a method that excites a pulse wave directly on the transmit antenna. A pulse of very short duration can be viewed as being a sum of sine waves of various frequencies and phases from low to high frequency, just like the spark discharge that Hertz used in his experiments.

Because short pulses can cover a very wide bandwidth, a UWB signal generates electromagnetic waves from 3 to 10 GHz. Imagine placing a sensitive narrowband receiver close to a transmitting UWB antenna. Turn it on and we hear that the UWB signal sounds like noise. Tune the receiver from 3 to 10 GHz and note how the average signal strength changes. A plot of the signal strength across the entire bandwidth of the UWB signal is called the *spectrum* of the signal.

The narrower the pulse width is, the wider the spectrum. The especially narrow pulse shown in Figure 8.1(e) has an extremely short interval between rising and falling edges. This means the spectrum is very wide. If we mathematically take the width to zero, but increase the voltage so that the total power is the same, we have the delta function that the physicist Dirac invented. This

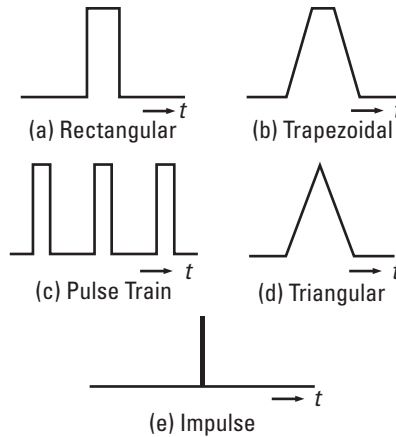


Figure 8.1 (a–e) Pulse waveforms.

function has the ultimate wideband spectrum, because it is a continuous, flat spectrum of constant amplitude at all frequencies. This is so extreme, however, that it can never be realized. For practical purposes, a narrow UWB pulse can be imagined having the spectrum of a Dirac delta function over a large band. Actual signals are not of infinite height and zero width like a delta function, so a real spectrum always tapers off at some point.

A Gaussian monocycle (Figure 8.2) is one common USB pulse waveform. Figure 8.3 shows the spectrum of pulses with two different pulse widths. The spectrum spreads out to higher frequencies when the width of the monocycle is cut in half from 600 to 300 ps. However, it is not a flat spectrum like the Dirac delta function.

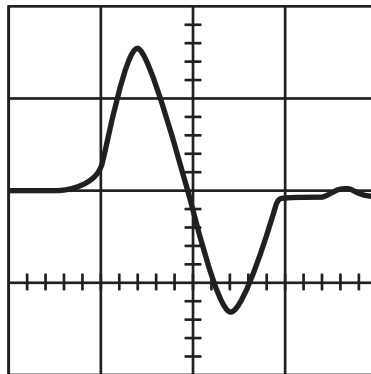


Figure 8.2 Pulse signal waves utilized in UWB.

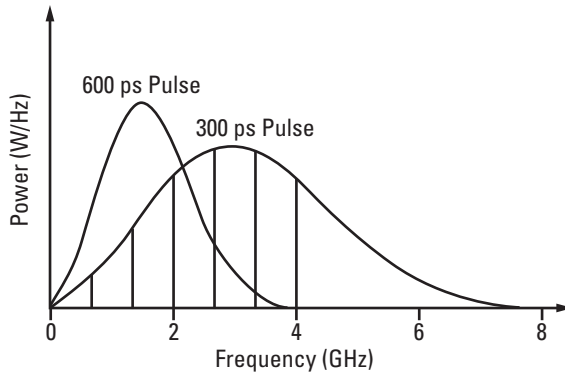


Figure 8.3 The spectrum of pulses of two different widths.

Digressing for just a moment, a picosecond (ps) is 1×10^{-12} seconds. To give a physical idea of how small a picosecond is, imagine taking a signal that is 1 second long and stretching it out between New York City and Los Angeles. A 300-ps pulse is then 1.2 mm long.

Returning to our pulse, the center frequency is obtained from the pulse width, τ , by following expression, which we calculate for a 600-ps pulse.

$$f = 1/\tau = 1/600 \times 10^{-12} = 1.7 \text{ GHz}$$

8.1.2 Log-Periodic Antennas

We start this section with a description of the log-periodic antenna because it is so well known and is easily designed. However, keep in mind that this antenna can also be quite large. In fact, if it is large enough that the pulse being transmitted takes more than a small fraction of a pulse width to travel the length of the antenna, then the antenna is not suitable for such a pulse transmission.

Figure 8.4 shows the structure of a log-periodic antenna. It has different length dipoles whose spacings and lengths are determined by means of the logarithmic function. Each element length and element spacing are set by the following expression.

$$L_{n+1}/L_n = S_{n+1}/S_n = k$$

Here, L_n is the length of the n th element, S_n is the spacing between n th and $(n + 1)$ th element and k is a scaling constant selected by the designer. The longest element is set to be one-half wavelength long at the lowest frequency and the shortest element is set to be one-half wavelength long at the highest

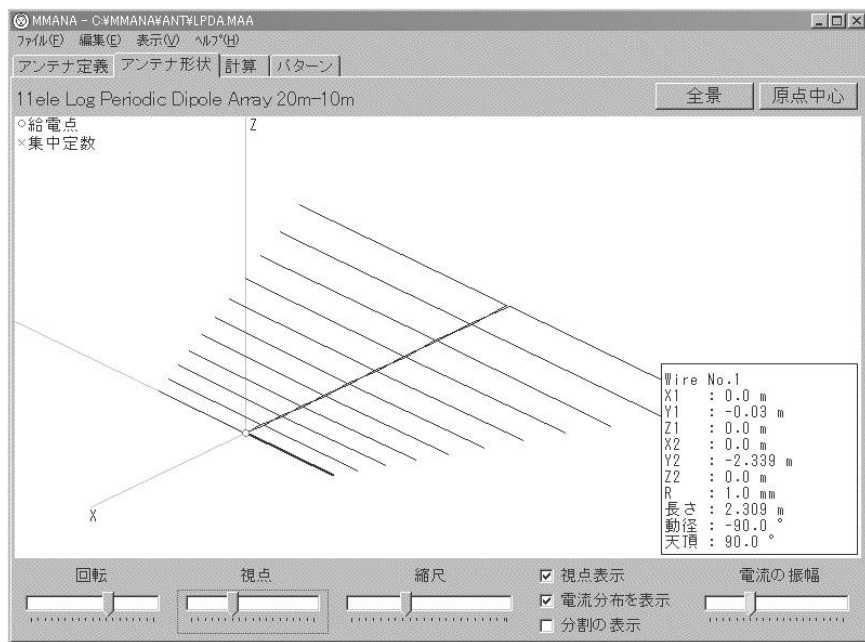


Figure 8.4 A log-periodic antenna model in the free MMANA software for 14–28 MHz.

frequency. There are published design curves for selecting k , and the apex angle, α , according to the desired gain and bandwidth. The apex angle is the angle between the two lines formed by the ends of the dipoles as they taper down to shorter lengths.

8.1.3 Design Example of a Log-Periodic Antenna

Here we provide a brief summary of a log-periodic design. Nearly all antenna text books already have good descriptions of this straightforward process. In addition, there are many free software tools available that automate the process. In this illustration, we wish to show typical numbers that can result from the process, so we describe a log-periodic dipole antenna design for UWB, which covers 3 to 10 GHz.

Given a desired gain of around 6.5 dBi, we look up the design curves (available in any of the popular antenna design books) and select the apex angle, α , to be 20° and k to be 1.27. We will include some design error margin and target a frequency ratio (bandwidth) of 4:1. The equation for taking a good guess at the number of required elements is next.

$$n = \ln(F)/\ln(k) = \ln(4)/\ln(1.27) = 1.386/0.239 = 5.8$$

Just to be safe, we round up and add two more. Our example uses nine elements. We design the first element to resonate at 15 GHz ($\lambda = 20$ mm). Thus, $L_1 = 10$ mm and the other elements are determined as follows:

$$L_2 = 10.0 \times 1.27 = 12.7 \text{ mm (15 GHz)}$$

$$L_3 = 12.7 \times 1.27 = 16.1 \text{ mm}$$

$$L_4 = 16.1 \times 1.27 = 20.5 \text{ mm}$$

$$L_5 = 20.5 \times 1.27 = 26.0 \text{ mm}$$

$$L_6 = 26.0 \times 1.27 = 33.0 \text{ mm}$$

$$L_7 = 33.0 \times 1.27 = 42.0 \text{ mm}$$

$$L_8 = 42.0 \times 1.27 = 53.3 \text{ mm}$$

$$L_9 = 53.3 \times 1.27 = 67.7 \text{ mm (2.2 GHz)}$$

Thus, a log-periodic antenna in free space is assumed and its maximum element length is as long as approximately 70 mm. In a similar manner, the initial spacing is read from the design curves to be 0.14λ , or 2.8 mm, and the other spacings are determined in the same manner as we determined the element lengths above.

When building the array, connect all the dipoles in parallel, one after another. Connect the feed line to the shortest element. Then cross the feed lines over to connect the first element to the second. Why is this? That is so there won't be a parallel resonant mode in between the resonant frequencies of the first and second dipole, just like the quasi-log-periodic broadband dipole in Section 6.2.4. As you consider a log-periodic antenna for your application, keep in mind that at any given frequency, only a small portion of the total number of dipoles are actually radiating power. Only dipoles that are near resonance are radiating.

The above design is for a log-periodic antenna in free space. To design a practical UWB antenna, we must additionally take into account things like a housing or case. If we are building it on a substrate, the dipole elements must be planar and include the wavelength shortening effect of the substrate.

8.1.4 Self-Complementary Antennas

A log-periodic dipole antenna on a substrate that might be better for a given application is the self-complementary structure discovered by honorary professor Yasuto Mushiaki at Tohoku University (Figure 8.5).

Notice that the pattern of conductor in the upper half is the same as the pattern of slots in lower half. This is what is meant by self-complimentary. The feed point is located at the left end, and the size of the ground (bottom plate) of substrate is 90×40 mm.

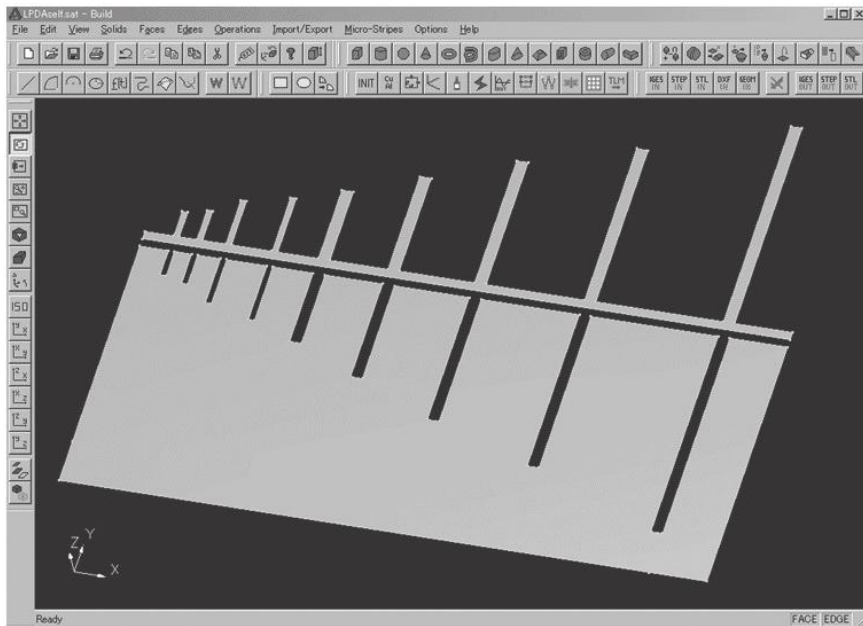


Figure 8.5 A self-complementary log-periodic dipole antenna.

For the purposes of simple illustration, the dimensions are taken from the design example above. There is no dielectric substrate so the size in free space is determined without the wavelength-shortening effect.

Figure 8.6 is a plot of the reflection coefficient of this antenna. It has a good low reflection response in the range of 3 to 10 GHz, with a little room for improvement at around 9 GHz. As the plot ripples along, you can imagine the resonating region moving within the antenna as we increase frequency.

The log-periodic dipole antenna covers a very wide band and is also used for receiving noise for EMC, electromagnetic compatibility, and testing. EMC testing involves listening for noise from electrical equipment as well as checking equipment for problems caused by electromagnetic radiation from other equipment.

Keep in mind that the log-periodic can have strong directivity, like the Yagi antenna. It is not suited for communication with a number of devices scattered in unknown locations, as is typical in wireless USB applications.

Figures 8.7 and 8.8 show the radiation patterns at 3 and 6 GHz. Directivity patterns differ slightly according to the frequencies but both patterns show strong directivity.

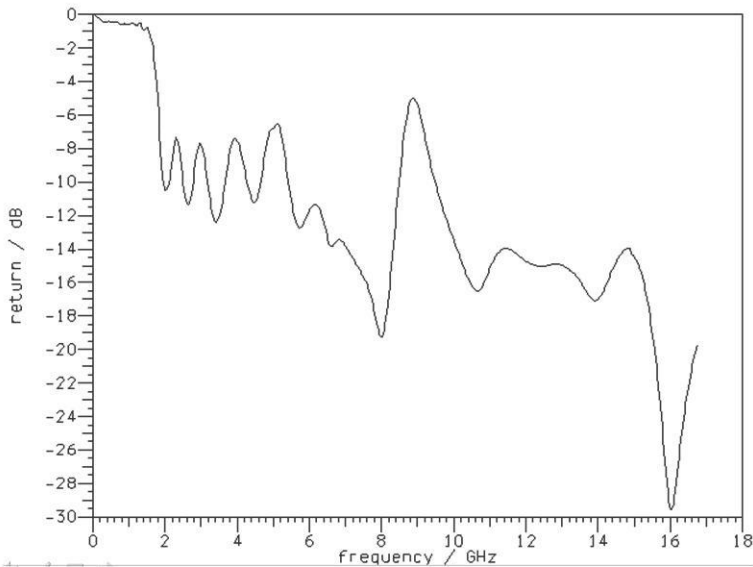


Figure 8.6 Reflection coefficient of the log-periodic dipole antenna shows low reflection from 3 to 10 GHz.

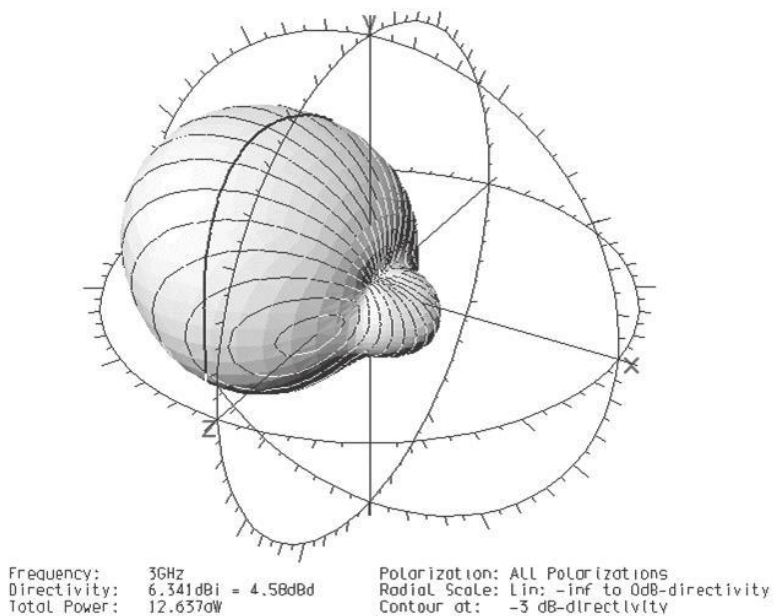


Figure 8.7 Log-periodic radiation pattern at 3 GHz.

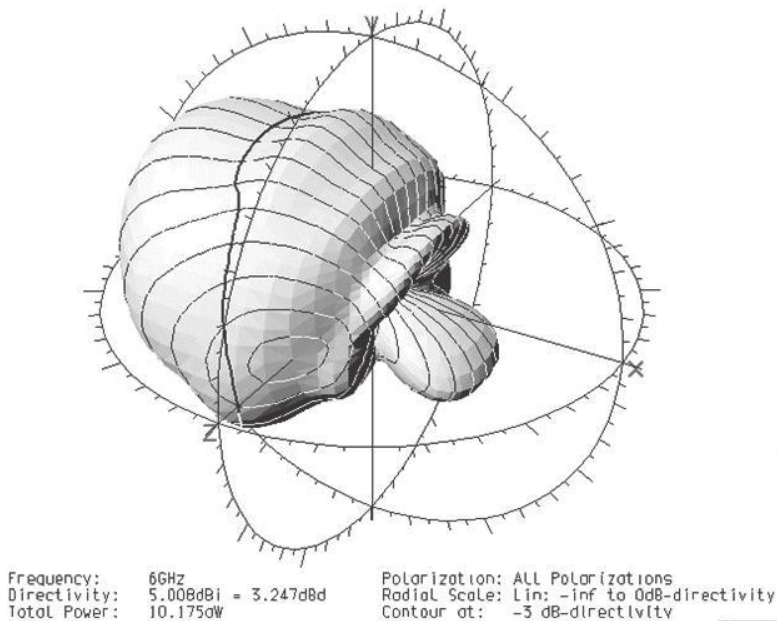


Figure 8.8 Log-periodic radiation pattern at 6 GHz.

The surface current distributions at 3 and 6 GHz are shown in Figures 8.9 and 8.10. We can see which elements are contributing to radiation at each frequency.

Considering the antenna operation for the entire band of a UWB system, the electromagnetic wave at 10 GHz is radiated from the short end of the antenna almost instantaneously when it arrives via the feed line. However, the electromagnetic wave at 3 GHz is radiated from the longer elements in the back. The signal from the feed line must travel all the way down the length of the antenna, then the radiated wave must travel all the way back to the front of the antenna. If the antenna size becomes large compared to the data signal pulse length, the shape of the transmitted pulse might be deformed.

8.1.5 Unbalanced Half-Trapezoid Dipole Antenna

A dipole antenna is a resonant antenna that uses standing waves, so without special consideration it is usable only for narrowband applications. Among nonresonant-type antennas, there are traveling wave antennas such as the three dimensional biconical antenna we saw in Chapter 5, and the bow tie antenna is a planar two-dimensional version.

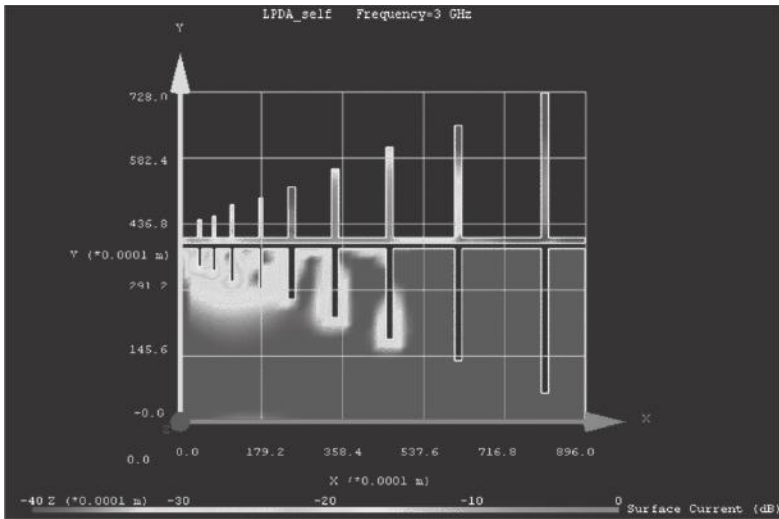


Figure 8.9 Log-periodic surface current distribution at 3 GHz.

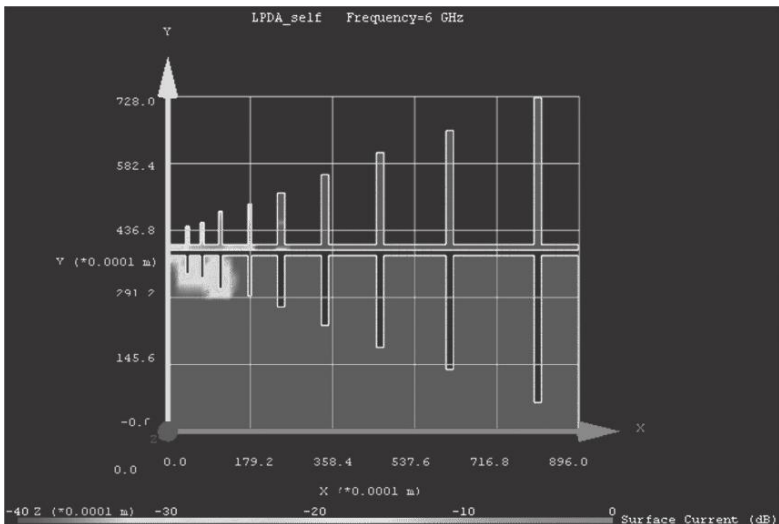


Figure 8.10 Log-periodic surface current distribution at 6 GHz.

Figure 8.11 is a Sonnet model that shows the surface current distribution on a bow tie antenna. In order to radiate a traveling wave smoothly, the antenna must be a few wavelengths long. However, this can make it too large for typical

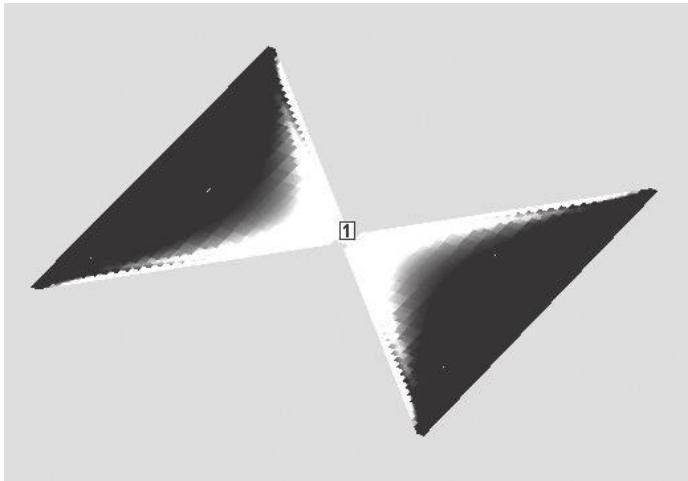


Figure 8.11 Surface current distribution on the bow tie antenna

home wireless appliances. Let's consider how we can reduce the size of the bow tie antenna as much as possible.

At the low frequency limit, the diagonal edges of the bow tie antenna are about one-half wavelength (Figure 8.11). At this frequency, the bow tie is similar to an ordinary one-half wavelength dipole antenna.

In Chapter 5, we used the Box sidewall as a ground for the triangle antenna in Sonnet simulation. However, in practical antennas, we use a finite area ground conductor, and when changing the size and shape of the ground conductor, the characteristics of a triangle antenna also change.

As we change the size and shape of the ground, if it becomes the same as the triangle element we return to the original bow tie antenna. So, why not intentionally design the elements asymmetrically? Figure 8.12 is an unbalanced half-trapezoid dipole antenna using this idea for wideband performance.

As we discussed at the end of Chapter 7, it is difficult to optimize multiple factors, simultaneously minimizing size and maximizing bandwidth. In an attempt to push those limits as far as possible, the built-in UWB antennas shown in Figure 8.12 were developed.

Figure 8.12(b) is an improved version of the antenna in Figure 8.12(a). Both are characterized in size and shape. The upper element and the lower element are different with the intent to realize broadband performance.

When looking at the VSWR plot in Figure 8.13, we see that the half-trapezoid dipole has improved performance around 9 GHz as compared with the trapezoid dipole. Figure 8.14(a) shows the surface current distribution at 3.2 GHz. We see strong current on both the edge of the semicircle and the edge

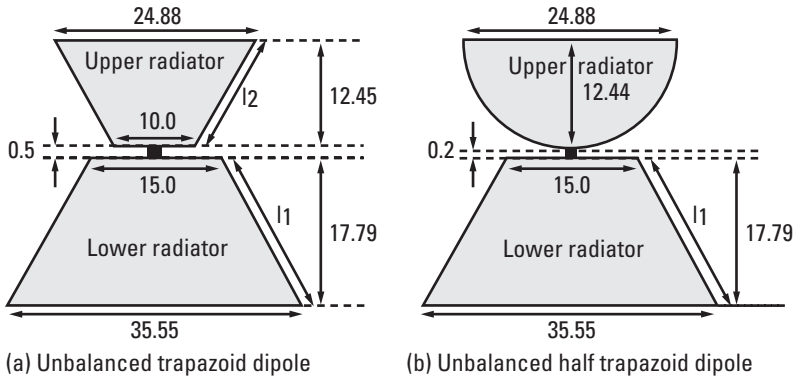


Figure 8.12 (a, b) Built-in UWB antennas; the upper and lower elements are different in size and shape.

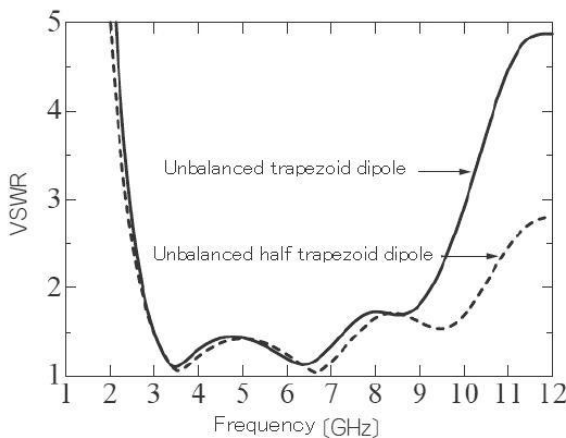


Figure 8.13 VSWR plot of the unbalanced half-trapezoid dipole antenna.

of the trapezoid as indicated by the arrows. Figure 8.14(b) at 6.6 GHz shows that the trapezoid is the main contributor to radiation. Figure 8.14(c) shows that at 10 GHz, the semicircle is most important.

When observing these three typical frequencies, we can see how it is possible to realize an ultra-broadband antenna with different portions of the antenna contributing to radiation at different frequencies.

For more information on the first antenna, see Horita Atsushi and Iwasaki Hisao, “Planar Trapezoid Dipole Antenna with Wideband Characteristics for UWB,” *IEICE Technical Report*, March 2005. For the second antenna, see

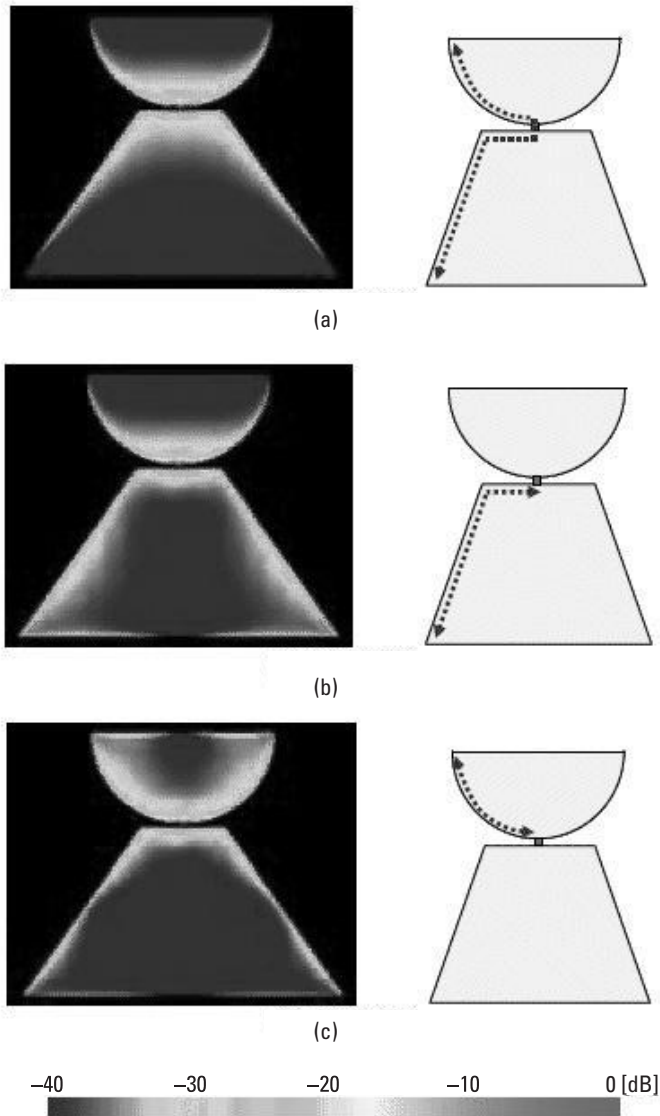


Figure 8.14 Surface current distribution on elements. (a) 3.2 GHz, (b) 6.6 GHz, and (c) 10 GHz.

Koshiji Fukuro, Eguchi Toshiya, and Sato Koichi, "Investigation of Radiators on Small Planar Antennas with Wideband Characteristics for UWB," Japan Institute of Electronics Packaging, November 2005.

8.2 Receiving Antennas for Digital Terrestrial Television

We now turn from broadband antennas to television antennas.

8.2.1 Dipole Antenna with Mesh Elements

The Japanese digital terrestrial television channels are assigned from 470 to 700 MHz. The dipole antenna with meshed elements that we designed in Chapter 7 (Figure 8.15) almost covers this band. If we adopt the idea of a dipole antenna with different length elements that we used in Chapter 6, perhaps we can cover a wider band still.

Figure 8.16 is a model that the first author improved in simulation based, with permission, on a receiving antenna for digital terrestrial television (which is covered by a registered Japanese design patent) that Mr. Shigekazu Shibuya, an authority on microwave propagation, has produced.

These are both broadband dipole antennas for receiving. Mr. Shibuya's antenna system has a metallic plate reflector in back. The separation between the plate and the antenna is optimally one-quarter of a wavelength at the center frequency, so it is designed with a metallic bent plate or metallic bent mesh (the dipole antenna is the original shape designed by Mr. Shibuya); see Figure 8.17. This results in a gain of more than 7.5 dBi at 600 MHz with a front-to-back ratio (F/B) of 10 dB.

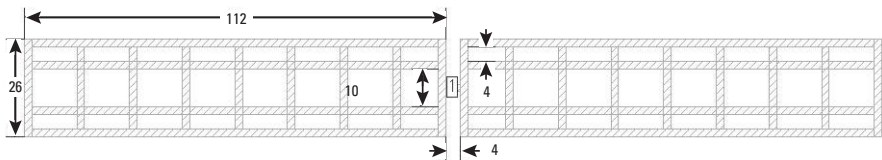


Figure 8.15 Dipole antenna with meshed elements.

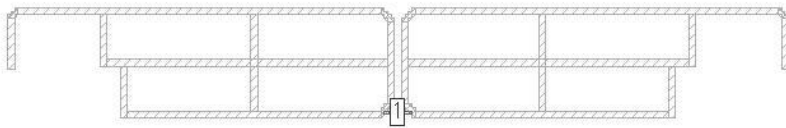


Figure 8.16 Improved version of a receiving antenna for digital terrestrial television.

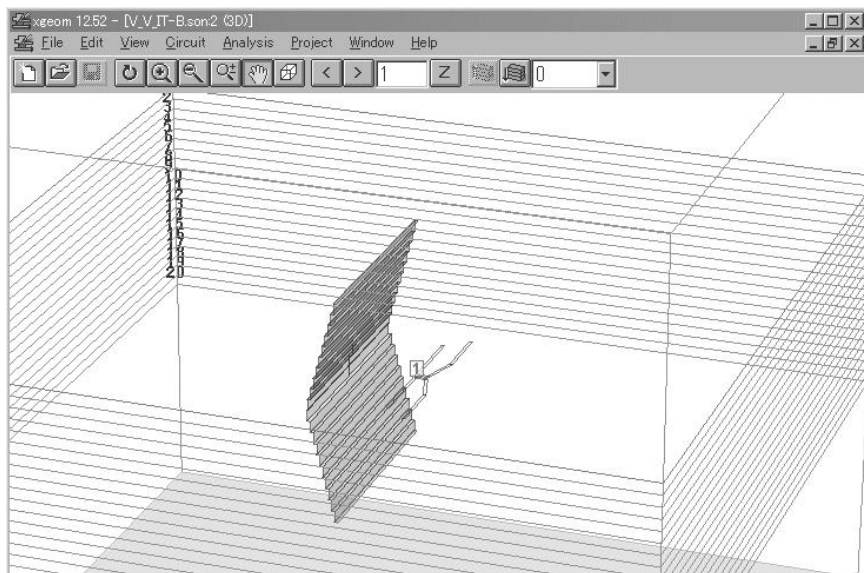


Figure 8.17 Antenna system design by Mr. Shibuya with a metallic plate as a reflector.

Figure 8.18 shows the VSWR of the antenna in Figure 8.16 (circles) and the VSWR of a model that has a reflector 140 mm to the rear (triangles). The

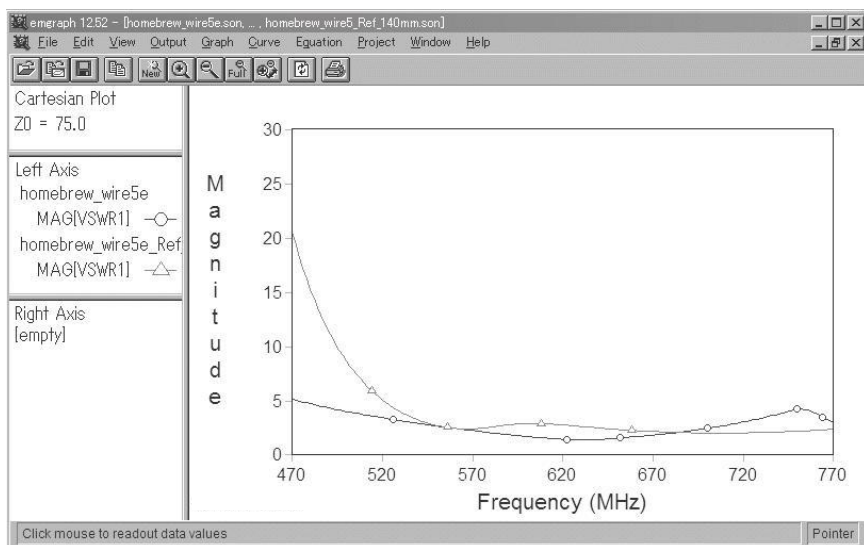


Figure 8.18 VSWR of the element in Figure 8.16 (circles) and the VSWR of a model with a reflector (triangles).

metallic plate increases the VSWR at lower frequencies. Fortunately, VSWR is not as important for receiving antennas as compared to transmitting antennas.

8.2.2 An Embedded Antenna for Receiving Digital Terrestrial Television

Cellular telephones that receive digital terrestrial television broadcasting used dipole and monopole antennas at first. However, having part of the antenna extending out of the body is undesirable. Thus, small embedded antennas were developed.

The first author, Hiroshi Kogure, made a presentation about the possibilities of small and embedded antennas, such as shown in Figure 8.19, at a seminar in 2003 (FPD International 2003), that covered early digital terrestrial television broadcasting.

Figure 8.20 shows the dimensions of this antenna, which has a substrate 2 mm thick with a relative permittivity of 9.8. The monopole antenna is bent several times, like an inverted L.

At the lowest digital terrestrial television frequency of 470 MHz, the wavelength is approximately 64 cm and a full size $\frac{1}{4} \lambda$ monopole 16 cm long. Because it is an element bent at three positions, the electromagnetic coupling with the ground conductor becomes strong and the antenna impedance decreases.

Figure 8.21 shows the impedance of this antenna. Looking at some typical frequencies, 400 MHz: $Z = 1.8 - j 37 \Omega$, 493 MHz: $Z = 3.6 + j 0 \Omega$, and 600 MHz: $Z = 29 + j 163 \Omega$. When directly connecting a 50Ω feed line, the mismatch reflection is large, so we need to design a matching network. After finding it by using a circuit simulator, as in Chapter 4, we can see that we use

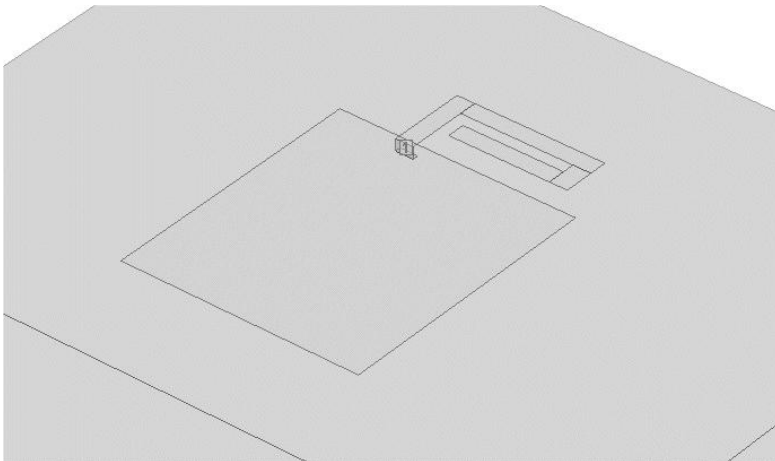


Figure 8.19 Small and embedded antenna for receiving digital terrestrial television.

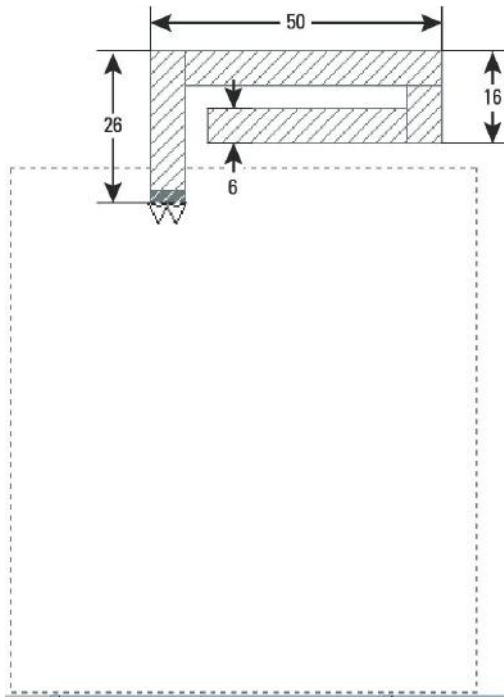


Figure 8.20 Dimensions of the miniaturized digital terrestrial television antenna.

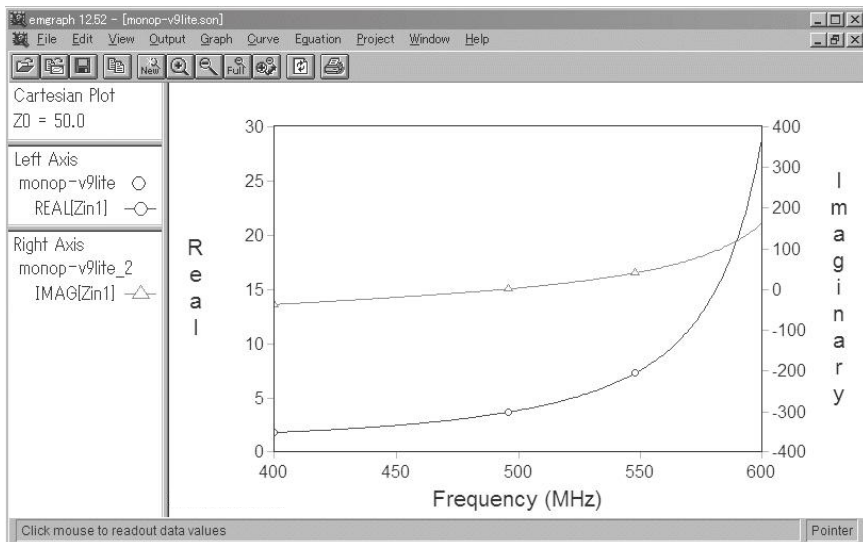


Figure 8.21 Impedance of the miniaturized digital terrestrial television antenna.

one L and one C together for any single frequency. However, this tends to be very narrowband. In order to tune the antenna over a broadband, variable inductors and capacitors are needed.

Because an ultramicrovariable inductor is not practical, we have found a method that varies capacitors only. An example of this method that was proposed at a seminar in 2003 uses variable capacitive diodes (varicaps) (Figure 8.22).

Figure 8.23 shows typical values for the variable capacitive diodes for three frequencies. The capacitance of each one is controlled by an applied DC voltage. It is ideal for matching because it can vary continuously, but the DC voltage must be maintained on the varicaps while in use.

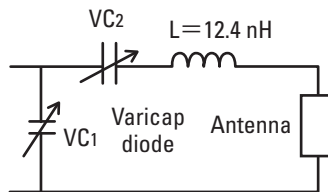


Figure 8.22 Matching network using variable capacitive diodes.

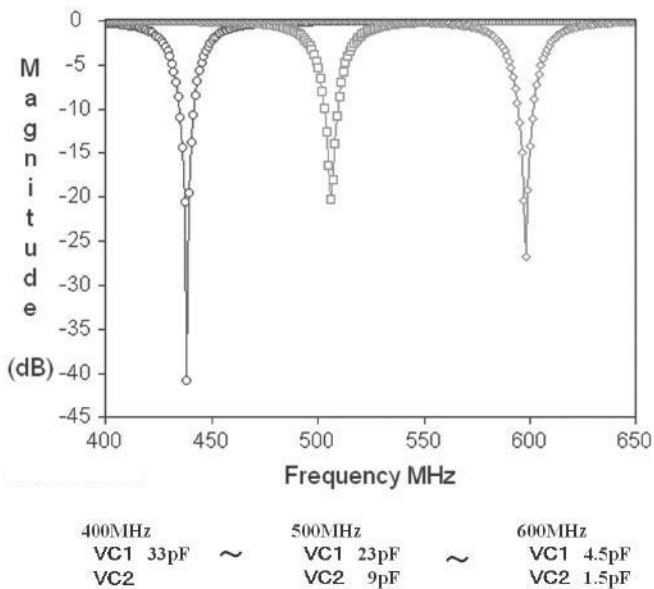


Figure 8.23 Values of VC1 and VC2, at three typical frequencies.

We could use an energy-saving method to switch various predetermined combinations of L s and C s for multiple bands in steps, without continuous adjustment. In receiving digital terrestrial television, the receiver might be randomly moved and reoriented so a nearly omnidirectional antenna pattern is desirable. Figure 8.24 shows the radiation pattern of this antenna. It is close to omnidirectional.

8.3 Antennas for Cellular Phones

Now we turn to one of the most important areas for compact antennas, cell phones.

8.3.1 Meander Line Monopole Antenna

The transmit/receive (transceive, or T/R) module for WiMAX introduced in Chapter 1 has an antenna at the end of a mass of printed wiring. A meander line is also sometimes used for cell phone antennas.

Figure 8.25 is a monopole antenna with a meander line element. It uses the Box sidewall of Sonnet as a ground, so it is a grounded monopole antenna. In an actual cell phone, a portion of the antenna might extend upwards from

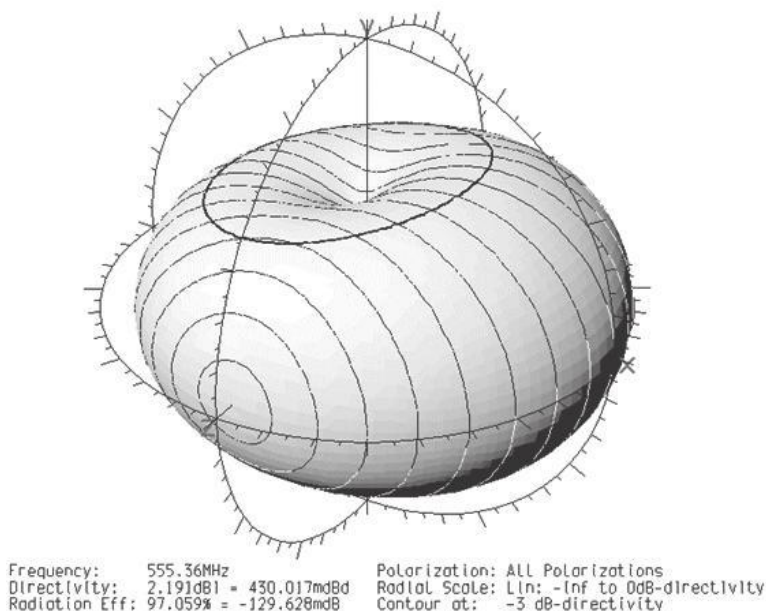


Figure 8.24 Radiation pattern of the miniaturized digital terrestrial television antenna.

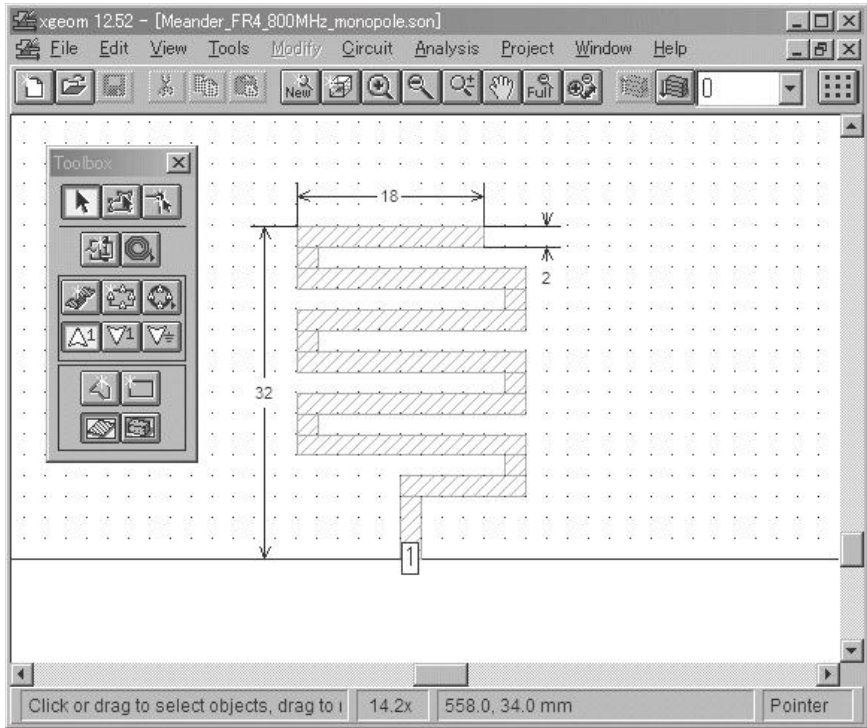


Figure 8.25 Model of a monopole antenna with a meander line element.

the housing. In this case, we need to include a ground plate of some sort to represent the body of the cell phone (see the next section). The box size is set to 1,024 mm for the x -direction and 512 mm for y -direction.

Figure 8.26 is a plot of the reflection coefficient and it indicates 3 dB at the resonant frequency of 800 MHz, suggesting that the value of R is not particularly close to 50Ω . The graph in Figure 8.27 is the input impedance. We see that at the resonant frequency, where X is zero, R takes on the small value of 8.4Ω . If this antenna is fed with 50Ω coaxial cable, the reflection is large and we need a matching circuit.

8.3.2 Meander Line Monopole Antenna with Ground Conductor

Figure 8.28 is a model with a 40×42 mm ground patch (representing the cell phone body) on a substrate. We will see if the monopole antenna still works. Figure 8.29 is a plot of the reflection coefficient of this antenna. It suggests that the resonance is at around 870 MHz. As the element length in monopole sec-

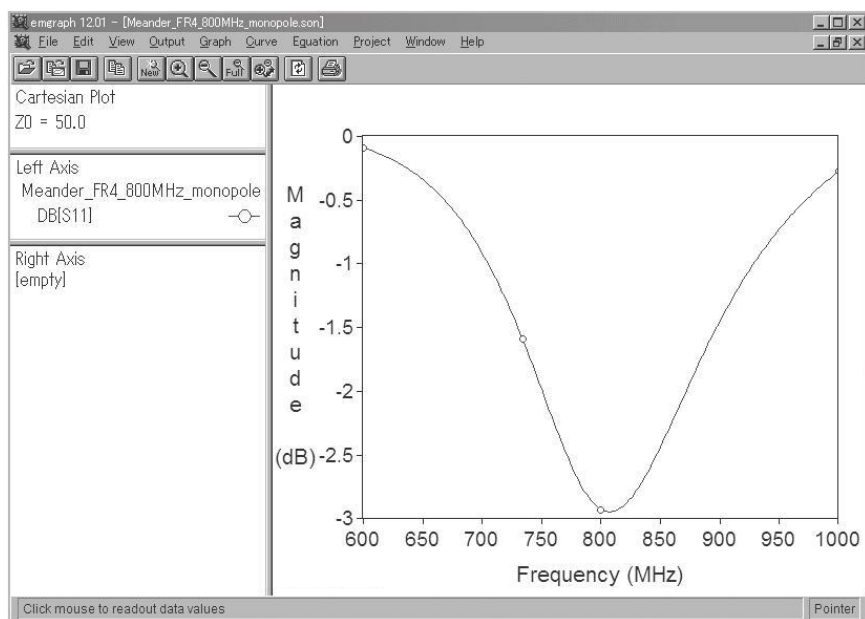


Figure 8.26 Reflection coefficient of the meandered monopole; the resonant frequency is around 800 MHz.

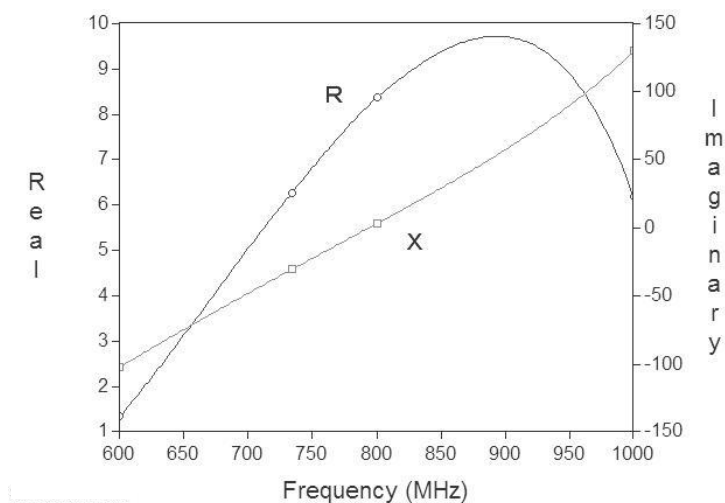


Figure 8.27 Input impedance of the meander line monopole antenna.

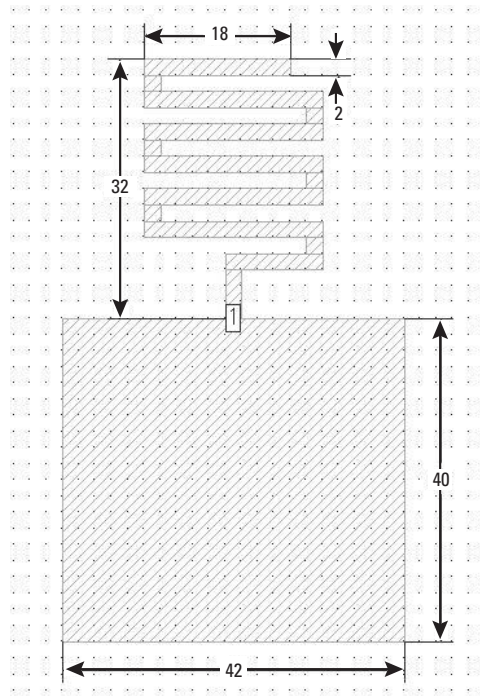


Figure 8.28 Monopole model with a ground patch on a substrate, roughly estimating what might happen in a real cell phone.

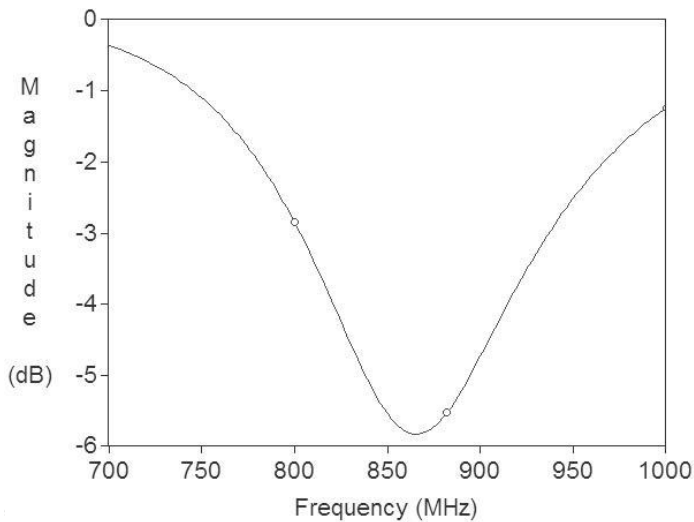


Figure 8.29 Reflection coefficient of monopole plus ground patch antenna.

tion is still the same, why does the resonance now shift to higher a frequency? Perhaps if we look at the current on the ground conductor we can see what is happening.

Figure 8.30 shows the surface current distribution at the resonant frequency and the strong current flow along the edge of the ground. Note that there is little current inside the ground patch. We know that the total amount of current flowing into (or out of) the ground patch must equal the total amount of current flowing out of (or into) the monopole antenna.

However, as shown in Figure 8.31, the areas of high-charge density are at the four corners of the ground patch and it looks just like the charge distribution on a capacitor plate. When the dimensions of ground patch are close to resonant dimensions, as we have here, it is no longer a ground, but rather a portion of a dipole antenna. It is just that one-half of the dipole has a wildly different shape from the other half. In this case we can see that this shape tends to accumulate charges, as shown in Figure 8.31.

Figure 8.32 shows the input impedance, which indicates that the reactance at the desired resonant frequency of 800 MHz (vertical axis to the right) is a negative -39Ω . Thus the input impedance is capacitive. Consequently, to add $+39\Omega$ of reactance, we make the meander element longer. This brings the resonance down to the desired frequency.

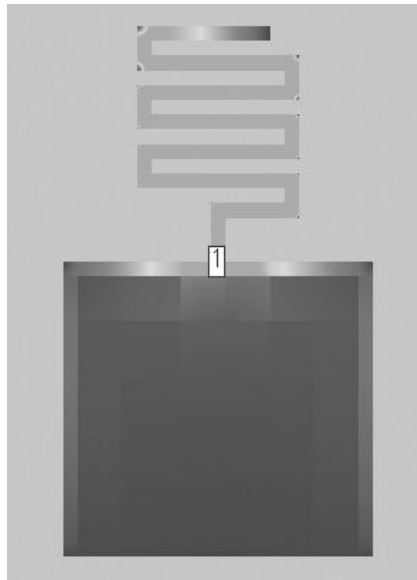


Figure 8.30 Surface current distribution for the monopole plus ground patch antenna.

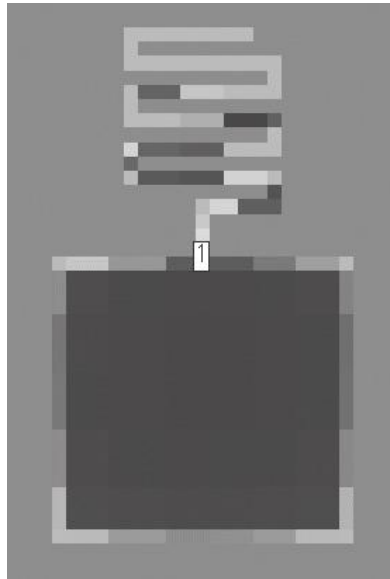


Figure 8.31 Charge distribution for the monopole plus ground patch antenna.

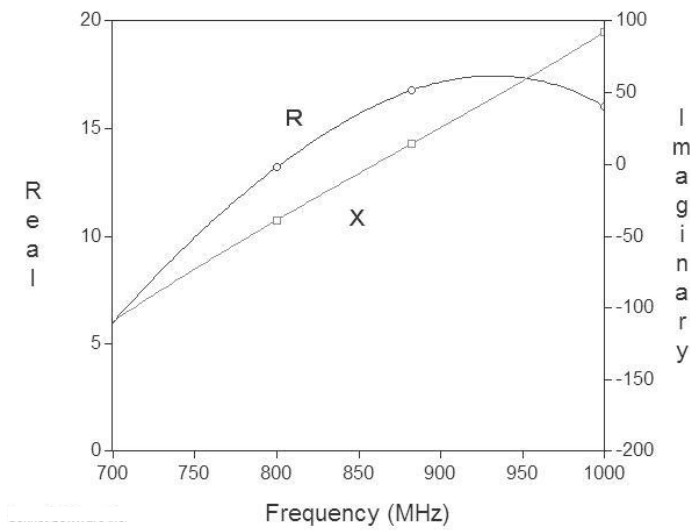


Figure 8.32 Input impedance of the monopole plus ground patch antenna.

8.3.3 Influence of Surrounding Metal Objects

Wireless portable terminals induce electromagnetic waves and energy directly by radiation from antennas. This radiation sometimes interferes with high-frequency circuits or high-speed circuits in its own housing. In Japan, this problem is called autotoxication. This is important in the field of signal integrity.

In order to prevent this problem, a metallic shield can be strategically placed near the small embedded antenna. However, such a metallic chassis itself induces electromagnetic energy and it can sometimes compromise antenna performance. To illustrate, we now simulate the effects of some shielding metal in the near field by means of a Sonnet box wall.

Figure 8.33 shows the dielectric layer settings for the model of a meander monopole antenna that has its own ground as simulated in previous section. The height of space under the substrate is changed to 5 mm. We also change the Box Bottom to metal to simulate a metallic shield or a chassis close to the antenna.

Figure 8.34 is the reflection coefficient, which indicates that the resonance has now shifted to around 740 MHz, and its reflection is as large as -0.43 dB. For the impedance, R is now approximately 1.2Ω , which is an extremely low value (Figure 8.35).

Sonnet does not display the current on the box bottom so we cannot view the current distribution there. However, this antenna is separated by only 5 mm and it seems reasonable that the current in the ground flowing around the feed point is strong. It is this coupling from the antenna to the ground plane that causes the input resistance to be so low.

8.3.4 Influence of Surrounding Dielectric Objects

In general, a portable wireless terminal is mounted in a plastic case. The proximity of the case to a small embedded antenna means that the resonant frequen-

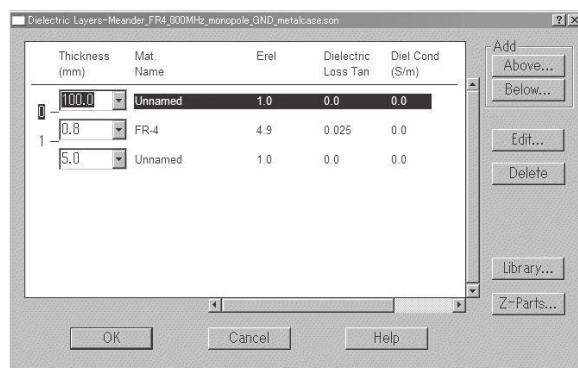


Figure 8.33 New dielectric layer settings for the meander monopole antenna.

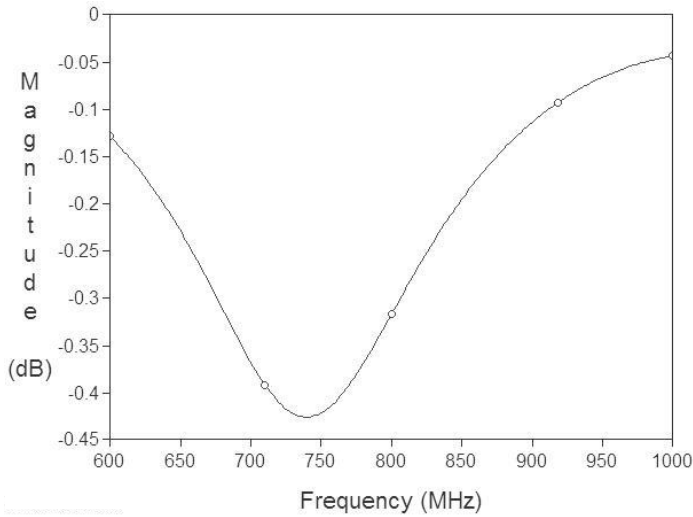


Figure 8.34 Reflection coefficient at 740 MHz is large.

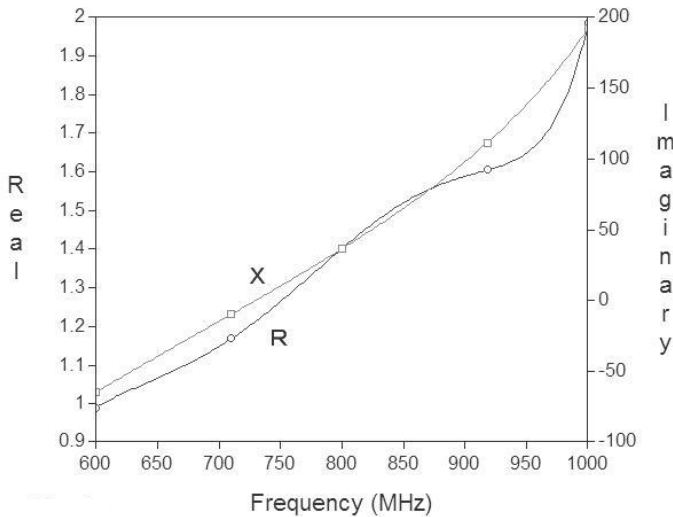


Figure 8.35 Impedance is very low; $R = 1.2\Omega$ at 740 MHz.

cy is shifted lower. Figure 8.36 shows a structure that includes an acrylonitrile butadiene styrene (ABS) resinous case 1 mm thick next to the substrate. Figure 8.37 shows the electrical attributes for ABS resin, with a relative permittivity of 3 and a $\tan\delta$ of 0.006.

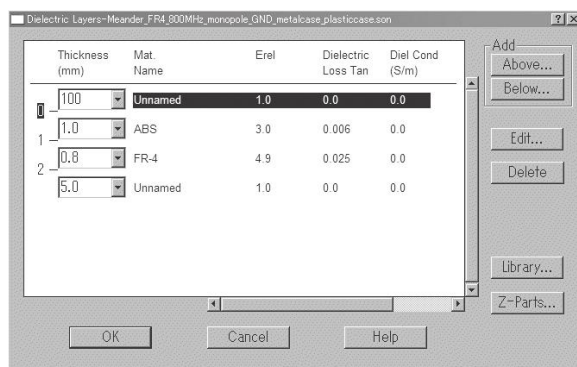


Figure 8.36 Setting the dielectric layers for an ABS resinous case 1 mm thick next to the antenna substrate.

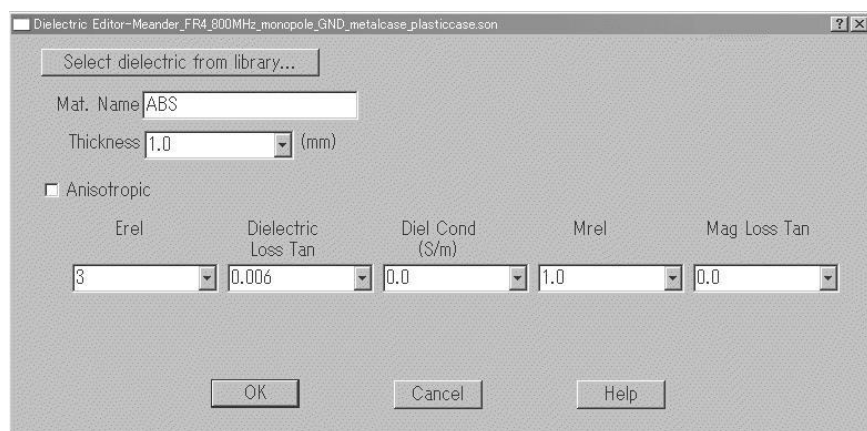


Figure 8.37 Electrical attributes for ABS resin.

Figure 8.38 is the reflection coefficient. We see that the resonance appears to shift from around 740 MHz down to around 700 MHz. The reflection coefficient is still around 0.4 dB. This is a large reflection. The impedance is shown in Figure 8.39, R is approximately 1.1Ω , still an extremely low value, so a matching circuit is needed when feeding with, say, a 50Ω source. The main effect of dielectrics near the antenna is to lower the resonant frequency.

8.3.5 Design of Matching Circuits

When the antenna input impedance is small, for example, 1Ω , a 50Ω source should not be connected as is. In the meander line monopole antenna, R was

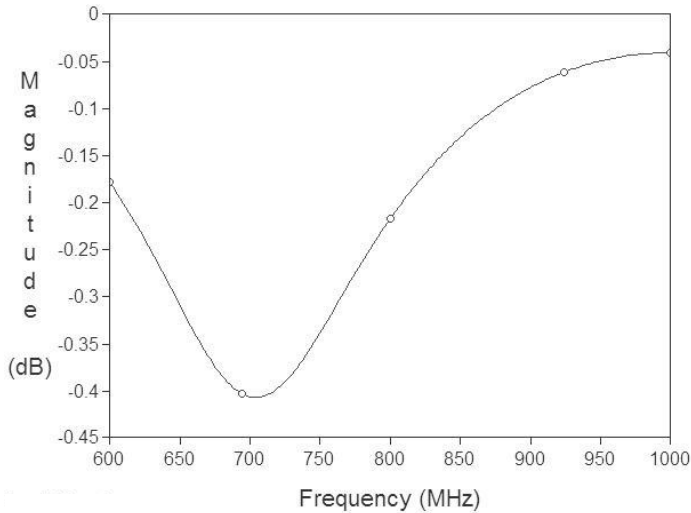


Figure 8.38 Reflection coefficient of the antenna including the ABS resin case.

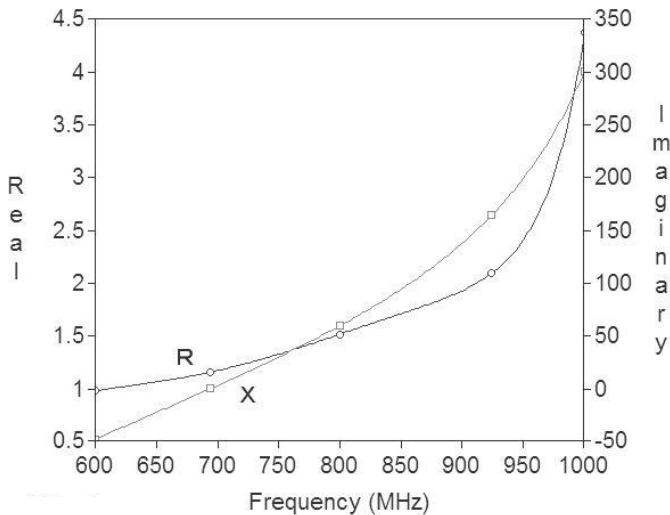


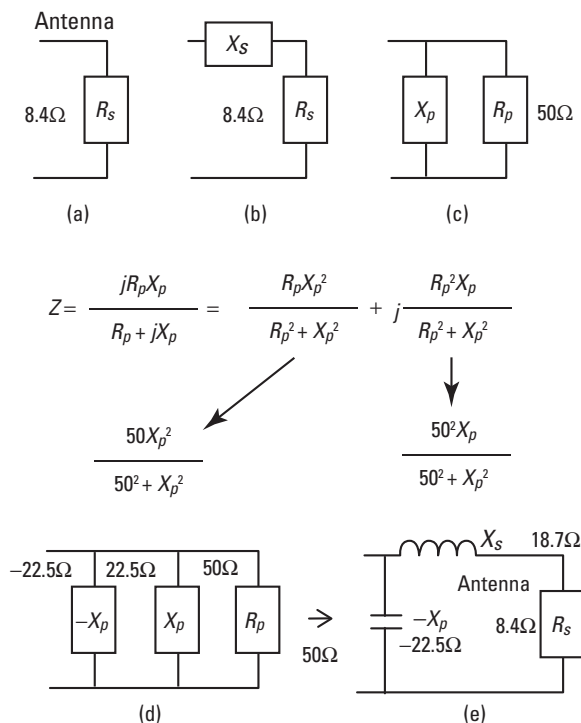
Figure 8.39 Input impedance of the antenna including the ABS case; R is 1.1Ω at 700 MHz.

8.4Ω at the resonant frequency of 800 MHz. Now we design a matching circuit and verify its performance using Sonnet Lite. In previous examples, we have found a matching network by using software made for that purpose. Microwave engineers can also find matching networks using a graphical aid called the

Smith chart, named after its inventor. Here, we show another method to find a matching network (Figure 8.40).

As shown in Figure 8.40(a), for illustration we assume that the real part, R_s , of the impedance at the resonant frequency of a miniaturized antenna is 8.4Ω . If the circuit connecting the reactance X_s in series as Figure 8.40(b) and the circuit connecting the reactance X_p in parallel as Figure 8.40(c) are equivalent, we can write the impedance Z of a parallel circuit by separating the real part and the imaginary part as shown in Figure 8.40 on the right.

The real part of this expression is equal to the real part R_s of Figure 8.40(b), 8.4Ω , then X_p is calculated to be 22.5Ω . Similarly, the imaginary part is equal to X_s of Figure 8.40(b). Then, by substituting $X_p = 22.5\Omega$, X_s is calculated to be 18.7Ω and the circuit of Figure 8.40(e) has the input impedance of 50Ω . Finally, we calculate the values L and C in Figure 8.41(e) for the resonant frequency of 800 MHz.



Real part = R_s of b (22.5Ω), then X_p is calculated to be 22.5Ω .

Imag part = X_s of (b), then using $X_p = 22.5\Omega$, $X_s = 18.7\Omega$.

Figure 8.40 (a–e) Procedure to find a matching circuit.

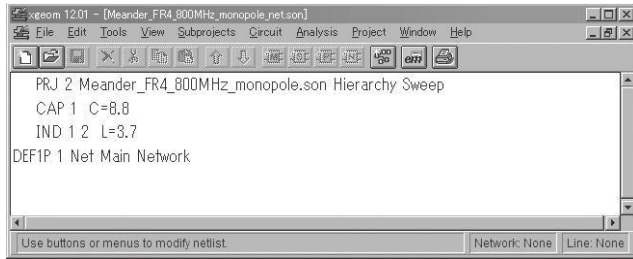


Figure 8.41 Netlist Project in Sonnet Lite.

$$L = \frac{18.7}{2\pi f} = \frac{18.7}{2\pi \times 800 \times 10^6} = 3.7 \text{ nH}$$

$$C = \frac{1}{22.5 \times 2\pi f} = \frac{1}{22.5 \times 2\pi \times 800 \times 10^6} = 8.8 \text{ pF}$$

Let's validate this calculation using a Netlist in Sonnet Lite. In the Sonnet Task Bar select Project->New Netlist, or in Sonnet xgeom select File->New Netlist. Then select Tools->Add Project Element, and assign the project file of the meander monopole antenna already simulated. Then, set Port Assignment so port 1 is connected to node 2 and the project file's ground is connected to GND.

Then select Tools->Add Modeled Element, and connect C and L to the appropriate nodes. The Port Assignment should end up as shown in Figure 8.41, with C connected between node 1 and GND, and L connected between nodes 1 and 2. The default is to create a 2-port, so, double-click on DEF2P and change it to a 1-port with port 1 connected to node 1. Node numbers, 1 and 2 in this example, are used as points to which we connect components when building a circuit as shown in Figure 8.42. Node 0 is the same as GND.

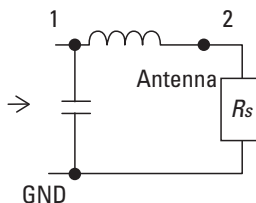


Figure 8.42 Matching circuit and node numbers.

Select Analysis->Setup and set Start to 600 MHz and Stop to 1,000 MHz in the Adaptive band sweep (ABS), and select Project->Analyze to start a simulation. Looking at the reflection coefficient seen from the matching network, the 50Ω normalized reflection is extremely low, but of course, narrowband, around 800 MHz (plot is omitted). Thus we can see that it is possible to connect a 50Ω coaxial cable directly.

The simple matching networks shown in our examples result from well-known techniques and are relatively easy to synthesize. However, as we have noted, the resulting match tends to be narrow band. One might think that we can get wider bandwidth if we were to use more lumped elements and a more complicated network. This is true up to a point; however, there is a limit to this approach. The limit is expressed formally and mathematically as *Fano's limit*. In practice, a lumped element broadband matching network for a resonant antenna, like the dipole, is not possible.

8.4 Small Antennas for Integrated Circuit Cards

Radio frequency identification (RFID) cards have seen tremendous growth over the last decade. Let's explore some practicalities of these credit-card-sized antennas.

8.4.1 Input Impedance of Integrated Circuits

The input impedance of integrated circuits (ICs) for RFID tags as seen by the antenna is not 50Ω , and it typically also includes reactance. Table 8.1 lists IC specs from the Web site of Impinj, Inc.

Table 8.1
IC Specs of MOMZA2™ from Impinj, Inc.

Configuration		Single-Ended	Differential	Shunt
Typical read power sensitivity		-11.5 dBm	-10.2 dBm	-10.5 dBm
Voltage sensitivity		190 mV _{RMS}	320 mV _{RMS}	180 mV _{RMS}
Linearized model of tag and mounting capacitance		$530\Omega \parallel 980 \text{ fF}$	$1050\Omega \parallel 680 \text{ fF}$	$380\Omega \parallel 1.87 \text{ pF}$
Recommended antenna impedance at minimum sensitivity	866 MHz (Europe)	$58 + j166\Omega$	$66 + j254W$	$24 + j92W$
	915 MHz (North America)	$52 + j158\Omega$	$59 + j242W$	$21 + j88W$
	956 MHz (Japan)	$48 + j153\Omega$	$55 + j233W$	$20 + j84W$

The impedance at 956 MHz (in the RFID band for Japan) is $48 + j153\Omega$; however when connecting a dipole antenna to this chip, we have to design the antenna's input impedance to be $48 - j153\Omega$. As just described, a dipole antenna is represented by an equivalent series resonant circuit, which has small impedance near resonance.

If you connect this antenna to the IC, the total reactance becomes zero and resistance R is matched at 48Ω for both. This is called conjugate matching and this method is typically used for RFID tags at UHF.

There is an entry of $530\Omega \parallel 980 \text{ fF}$ in the IC data shown in the third row of the second column headed "Single-ended" of Table 8.1. This is the equivalent parallel resistance and capacitance of the mounted tag before attaching the antenna. In UHF tags, there is an "antenna" type that uses magnetic coupling just like the microloops we use at 13.56 MHz. Impinj's button tag is one of these. In this case the button tag is represented by an equivalent parallel resonant circuit that resonates with a C of 980 fF. If the product sheet specifies only the series input impedance instead, $48 + j153$ ohms for example, simply perform a series to parallel conversion to get the parallel impedance with the expression shown in Figure 8.43.

8.4.2 Matching Method for IC Including Reactance

Let's assume the input impedance obtained after simulating an antenna is $20 + j225\Omega$, and the impedance of the IC is, for example, $80 + j100\Omega$. We will design a circuit using a circuit simulator (any of many available circuit simulators will do) to find a matching circuit. In Figure 8.44, these impedances are converted into values for the R and L circuit elements (when the reactance is negative, calculate the value of C). After drawing the circuit in a circuit simulator as in Figure 8.45, the circuit shown in Figure 8.46 is derived by using a typically available option to synthesize a matching circuit.

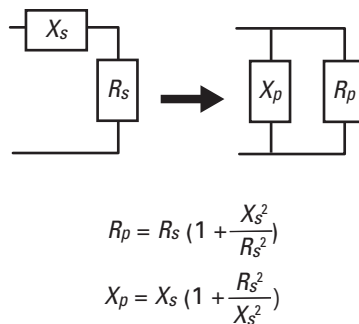


Figure 8.43 Series to parallel conversion equations.

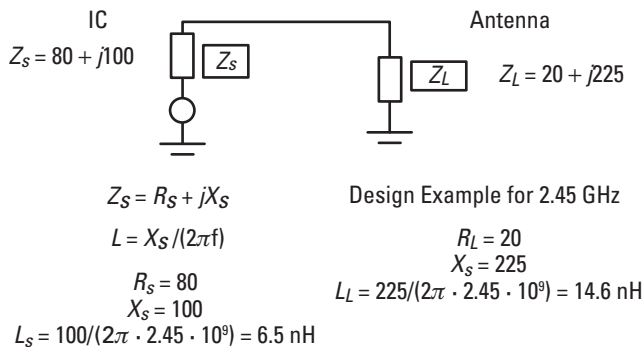


Figure 8.44 Impedances are converted into values of R and L of circuit elements.

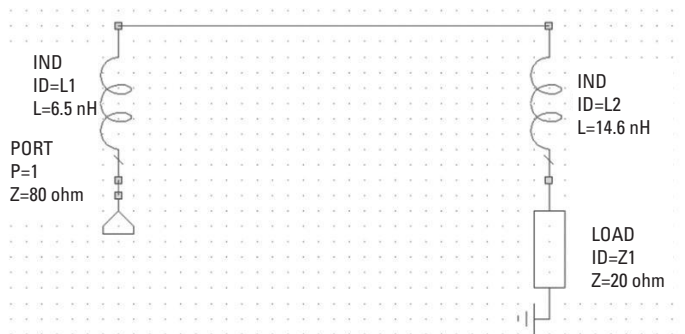


Figure 8.45 Drawing the circuit in a circuit simulator.

We obtain a matching circuit with a 0.4-pF capacitor in series and a 1.34-pF capacitor in shunt. In other cases, when the impedance at the resonant frequency is not 50Ω , this method improves the match significantly.

8.4.3 Changing the Shape of Dipole Elements to Achieve a Match

Next we describe a method for realizing a match by just modifying the shape of a dipole antenna for UHF tags. We illustrate the technique for the case when the IC impedance is $20 - j150\Omega$.

First, the input resistance is small, 20Ω , so we can use the fact that the input R of a dipole decreases when folding the elements. Then we adjust the shape and the dimensions of the antenna so that R drops to 20Ω as determined by using an electromagnetic field simulator.

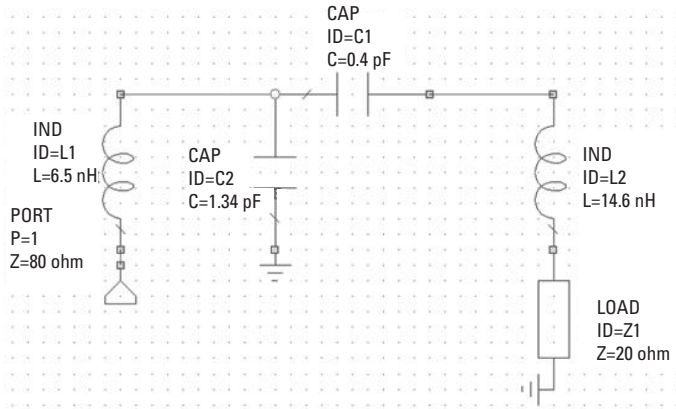


Figure 8.46 Derived circuit by using the feature to design a matching circuit.

The reactance X at the IC output is capacitive (negative). This allows us to build an inductive antenna (positive reactance). Because L of a dipole antenna increases when lengthening the elements, it would be nice to fold the elements as we lengthen them.

Figure 8.47 is the antenna designed using this procedure. Figure 8.48 is the input impedance and it is close to $20 + j150\Omega$. When the available increase in L is insufficient in this method, we can put a loop of wire in the vicinity of the IC as shown in Figure 8.49 to increase L .

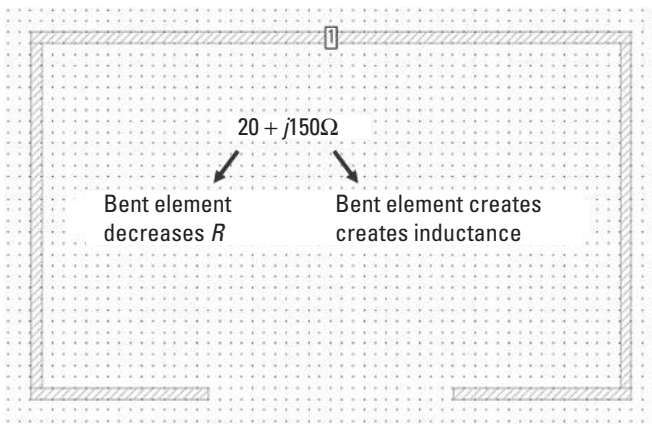


Figure 8.47 Bent dipole antenna.

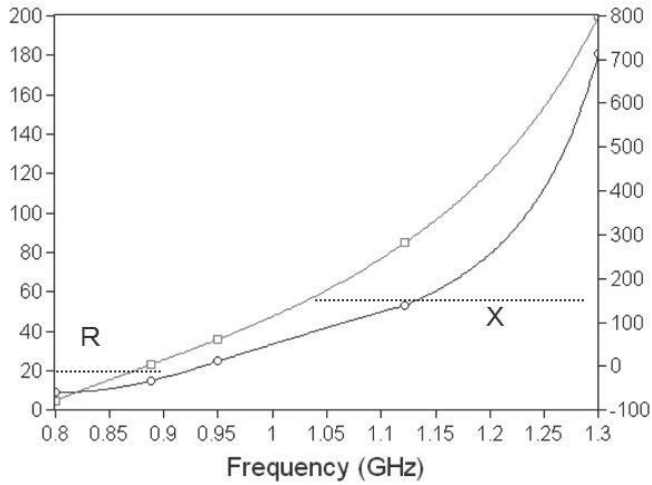


Figure 8.48 Input impedance of this antenna.

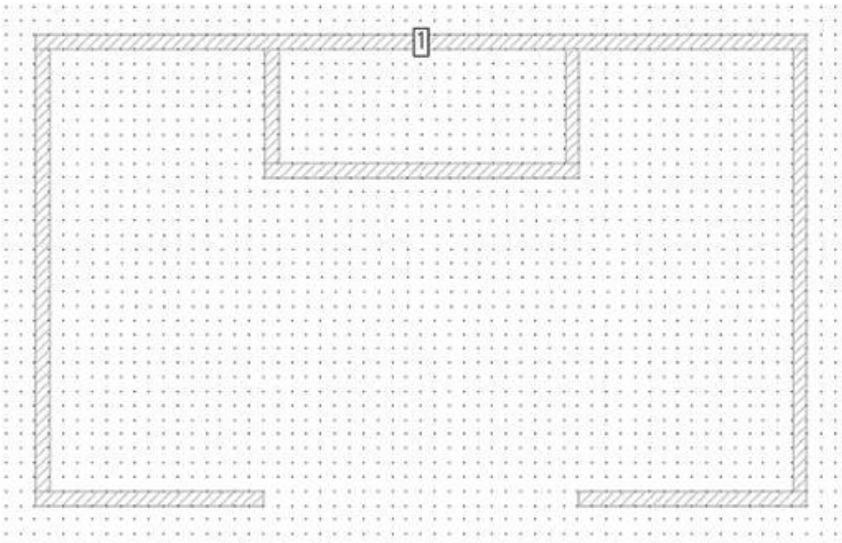


Figure 8.49 A loop in the vicinity of the IC increases L .

8.5 The Wireless World Is Expanding

Few areas have seen the growth in revenue and in technology that we have seen in all the various forms of wireless over the last decade. Let's go over where we are and where we might go.

8.5.1 Small Embedded Antennas Have a Bright Future

In this book, we have learned the fundamentals of antennas using various examples. As the wireless world blossomed, development of small antennas advanced rapidly. When we investigate small antennas, we learn about their performance under harsh design constraints, achieving specifications which cannot be realized by conventional and full-sized antennas.

The three semicylindrical objects shown in Figure 8.50 are antennas made of ceramics developed by Antenova Ltd. These are fabricated from high dielectric material to pull in nearby electric fields, and they resonate at a specific frequency. When placed on top of and in the vicinity of a microstrip line, strong electromagnetic fields couple electromagnetic energy into these dielectric antennas.

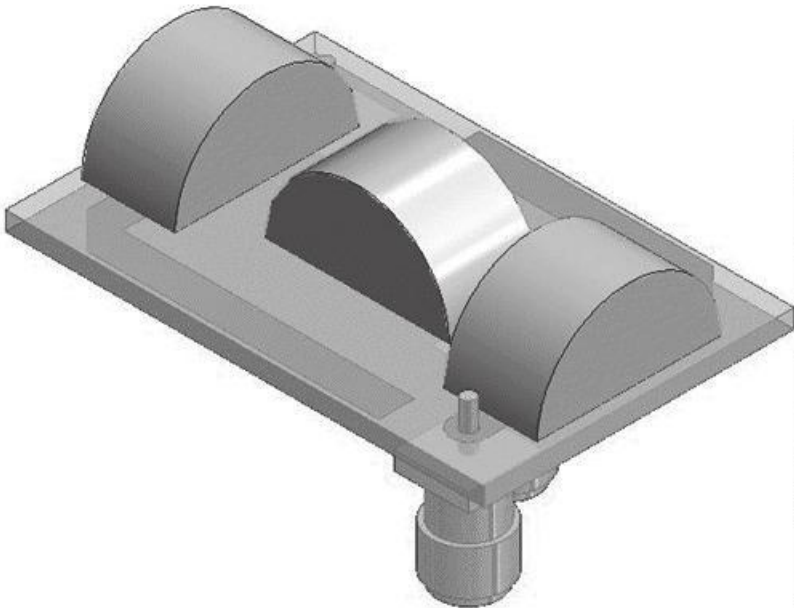


Figure 8.50 Three semicylindrical antennas made of ceramics.

The electric field vectors distribute along the curved surface of each antenna, and they are radiate efficiently into the space around the curved surface. As there are no conductors inside this antenna, its means of operation is entirely different from Hertz's original dipole antenna.

Like these antennas, more and more exotic antennas, which are typically not in the textbooks, are proliferating. Requirements for miniaturizing antennas that have never before existed inevitably leads to increasing use of electromagnetic field simulators for design and development.

In early days of Hertz, Nagaoka, and Marconi, they could experiment on any wavelength at will. However, today's wireless systems come into use at every frequency, with modern day frequency allocations nearly full. Despite the fact that those wavelengths can cover extremely wide ranges, like 7.5 km (40 kHz) for radio controlled clocks, 5 cm (5.8 GHz) for electronic toll collection systems (ETCs), and so forth, and they must all be palm-sized. This is the wireless world's great constraint. We, as antenna engineers, must accept this challenge.

Figure 8.51 is a microtag design that the first author, Hiroaki Kogure, helped improve. He redesigned this tag based on the method in Chapter 6. This is a tag for ready-mixed concrete to control the production record. It was produced by Sobu Ready-mixed Concrete, Co. The major diameter is 17 mm and the minor diameter is 12 mm. It has an embedded microantenna (coil), in spite of a 22m wavelength at 13.56 MHz.

As the tags are to be thrown in by the handful when mixing the materials of ready-mixed concrete, they are covered with a special plastic to maintain their integrity in high-moisture alkaline environments. They are designed so that the production district of the cement, composition of materials, and the name of the manufacturer are written into the tag using a reader-writer. In this case, once it is written, it is impossible to tamper with the information.

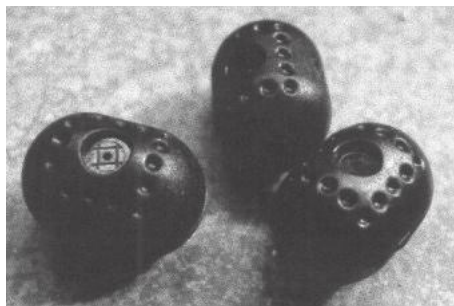


Figure 8.51 Microtags produced by Sobu Ready-mixed Concrete, Co.

In the systems involved in public infrastructures, reliability is a top priority. The wireless world extends more and more, and small antennas as typified by IC tags are a cutting edge technology that plays a role in the developing field of information and communication technology (ICT).

The future of small antennas and the wireless world is bright.

Appendix

Sonnet Lite™ Installation

You can install Sonnet Lite from the DVD that accompanies this book, or you can go to <http://sonnetsoftware.com/products/lite/> and download the latest version.

Before installing Sonnet Lite, please read the License Agreement. An important term in the license agreement is that Sonnet Lite cannot be used for publishing competitive comparisons; doing so violates the License Agreement.

Follow these steps to install Sonnet Lite:

1. To Install Sonnet Lite:
 - Download or copy from the DVD the following file to any directory:
sl1253.exe (50 Mbytes)
2. Download Adobe Acrobat® Reader® if you don't already have it.

You need Adobe Acrobat Reader 7.0.8 or above to read the manuals and tutorials. You will need the manuals to get the most out of Sonnet Lite. If you already have Adobe Acrobat Reader installed on your machine, you may skip this step. To download Adobe Acrobat Reader, go to <http://get.adobe.com/reader/>.
3. Sonnet Lite can be run by selecting Start->Programs->Sonnet 12.53->Sonnet. We highly recommend that you go through the tutorials, which can be found by selecting Getting Started from the Help menu. A little effort spent in the tutorials will save you time and make

you productive with Sonnet Lite in the quickest way possible. There are also numerous tutorials and videos on the Sonnet Web site.

4. Register Sonnet Lite by selecting Start->Programs->Sonnet 12.53->Register. You may use Sonnet Lite without the requirement to register the software, and solve problems that require up to 1 MB of RAM to analyze. However, if you register your copy of Sonnet Lite, we'll send you a license that will enable you to solve problems requiring up to 16 MB of RAM to solve. It only takes a few minutes to register your software.

If you experience any trouble installing or registering, please check the Sonnet Lite Troubleshooting Guide.

About the Authors

Hiroaki Kogure received a BSEE from Tokyo University of Science in 1977. In 1977, he joined Hitachi Engineering Co. Ltd., and was engaged in the development of electric power control systems and in the test production of a pioneer object oriented precompiler language “objC” in 1985. Since 1992, he has been a representative of Kogure Consulting Engineers.

Dr. Kogure completed his doctorate (Dr. Eng.) in 1998 in electromagnetic field analyses at Tokyo University of Science. He is also a part-time lecturer at the Tokyo University of Science since 2004 and Tokyo City University (Ex-Musashi Institute of Technology) since 2006.

He primarily works as a registered professional engineer of information technologies to support research and development for leading companies in Japan. Dr. Kogure has written many technical books in the fields of electromagnetics, high frequencies, EMC problems, and antennas.

Yoshie Kogure received a B.A. in Chinese literature from Waseda University in 1983. Since 1992, she has been supporting Kogure Consulting Engineers as a technical writer and a translator. She has written many technical books in collaboration with Dr. Kogure.

James C. Rautio received a BSEE from Cornell in 1978, an MS in Systems Engineering from University of Pennsylvania in 1982, and a Ph. D. in electrical engineering from Syracuse University in 1986. From 1978 to 1986, he worked for General Electric, first at the Valley Forge Space Division, then at the Syracuse Electronics Laboratory. At this time he developed microwave design and measurement software, and designed microwave circuits on Alumina and on GaAs. From 1986 to 1988, he was a visiting professor at Syracuse University and at Cornell. In 1988 he went full time with Sonnet Software, a company

he had founded in 1983. In 1995, Sonnet was listed on the Inc. 500 list of the fastest growing privately held US companies, the first microwave software company ever to be so listed. Dr. Rautio was elected a fellow of the IEEE in 2000 and received the IEEE MTT Microwave Application Award in 2001. He has lectured on the life of James Clerk Maxwell over 100 times.

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